IN THE UNITED STATES DISTRICT COURT
FOR THE DISTRICT OF DELAWARE

CIF LICENSING, LLC, d/b/a GE LICENSING,	)	
Plaintiff,	)	C.A. No. ______________
v.	)	**JURY TRIAL DEMANDED**
AGERE SYSTEMS INC.,	)	
Defendant.	)	

COMPLAINT FOR DAMAGES AND INJUNCTIVE RELIEF
(DEMAND FOR JURY TRIAL)

Plaintiff CIF Licensing, LLC, d/b/a GE Licensing ("GE Licensing"), by its attorneys, complains against defendant Agere Systems Inc. ("Agere"), and alleges as follows:

THE PARTIES

1. Plaintiff GE Licensing is a limited liability company organized and existing under the laws of the State of Delaware with its principal place of business in Princeton, New Jersey 08540.

2. On information and belief, Defendant Agere is a corporation organized under the laws of the State of Delaware, with its principal place of business at 1110 American Parkway N.E., Allentown, Pennsylvania 18109. Agere is in the business of making, selling, and importing modems and modem systems.

JURISDICTION AND VENUE

3. This is an action arising under the patent laws of the United States, 35 U.S.C. § 101, *et seq*. This Court has exclusive subject matter jurisdiction under 28 U.S.C. §§ 1331 and 1338(a).

4. Personal jurisdiction for this action is proper in this Court as defendant Agere is incorporated under the laws of the State of Delaware.

5. Venue is proper in this judicial district under 28 U.S.C. §§ 1391(b) and (c) and 1400(b) as defendant is incorporated under the laws of the State of Delaware.

THE PATENTS

6. United States Letters Patent No. 5,048,054 ("the '054 Patent"), entitled "Line Probing Modem" issued on September 10, 1991 to inventors Vedat M. Eyuboglu and Ping Dong. A copy of the '054 Patent is attached hereto as Exhibit A.

7. United States Letters Patent No. 5,428,641 ("the '641 Patent"), entitled "Device and Method for Utilizing Zero-Padding Constellation Switching with Frame Mapping" issued on June 27, 1995 to inventor Guozhu Long. A copy of the '641 Patent is attached hereto as Exhibit B.

8. United States Letters Patent No. 5,446,758 ("the '758 Patent"), entitled "Device and Method for Precoding," issued on August 29, 1995 to inventors Vedat M. Eyuboglu. A copy of the '758 Patent is attached hereto as Exhibit C.

9. United States Letters Patent No. 6,198,776 ("the '776 Patent"), entitled "Device and Method for Precoding Data Signals for PCM Transmission," issued on March 6, 2001, to inventors Vedat M. Eyuboglu, Pierre A. Humblet, and Dae-young Kim. A copy of the '766 Patent is attached hereto as Exhibit D.

10. GE Licensing is the owner of all rights, title and interest in and to the '054 Patent, the '641 Patent, the '758 Patent, and the '776 Patent ("the GE Patents"), by assignment, with the right to recover damage for all past and future infringement of the GE Patents.

BACKGROUND

11. The GE Patents cover inventions relating to hardware and software modems and modem systems.

12. On information and belief, Agere has imported into the United States, used, manufactured, sold and/or offered for sale in the United States, products covered by the GE Patents.

13. On information and belief, Agere has had actual and/or constructive notice and knowledge of the GE Patents. The filing of this Complaint also constitutes notice in accordance with 35 U.S.C. § 287. Despite such notice, Agere continues to import into the United States, use, manufacture, sell, and/or offer for sale in the United States products covered by the GE Patents.

COUNT I

14. GE Licensing repeats and realleges the allegations in paragraphs 1-13 as though fully set forth herein.

15. On information and belief, Agere has infringed, contributed to the infringement of, and/or induced infringement of the '054 Patent by making, using, selling, offering for sale, or importing into the United States, or by aiding and abetting others to make, use, import into, offer for sale, or sell in the United States, products that incorporate or practice the invention of the '054 Patent.

16. On information and belief, Agere's infringement of the '054 Patent has been willful. Agere's continued infringement of the '054 Patent has damaged and will continue to damage GE Licensing.

17. On information and belief, unless enjoined by this Court, Agere will continue to infringe the '054 Patent, and GE Licensing will continue to suffer irreparable

harm for which there is no adequate remedy at law. Accordingly, GE Licensing is entitled to permanent injunctive relief against such infringement pursuant to 35 U.S.C. § 283.

COUNT II

18. GE Licensing repeats and realleges the allegations in paragraphs 1-13 as though fully set forth herein.

19. On information and belief, Agere has infringed, contributed to the infringement of, and/or induced infringement of the '641 Patent by making, using, selling, offering for sale, or importing into the United States, or by aiding and abetting others to make, use, import into, offer for sale, or sell in the United States, products that incorporate or practice the invention of the '641 Patent.

20. On information and belief, Agere's infringement of the '641 Patent has been willful. Agere's continued infringement of the '641 Patent has damaged and will continue to damage GE Licensing.

21. On information and belief, unless enjoined by this Court, Agere will continue to infringe the '641 Patent, and GE Licensing will continue to suffer irreparable harm for which there is no adequate remedy at law. Accordingly, GE Licensing is entitled to permanent injunctive relief against such infringement pursuant to 35 U.S.C. § 283.

COUNT III

22. GE Licensing repeats and realleges the allegations in paragraphs 1-13 as though fully set forth herein.

23. On information and belief, Agere has infringed, contributed to the infringement of, and/or induced infringement of, the '758 Patent by making, using,

selling, offering for sale, or importing into the United States, or by aiding and abetting others to make, use, import into, offer for sale, or sell in the United States, products that incorporate or practice the invention of the '758 Patent.

24. On information and belief, Agere's infringement of the '758 Patent has been willful. Agere's continued infringement of the '758 Patent has damaged and will continue to damage GE Licensing.

25. On information and belief, unless enjoined by this Court, Agere will continue to infringe the '758 Patent, and GE Licensing will continue to suffer irreparable harm for which there is no adequate remedy at law. Accordingly, GE Licensing is entitled to permanent injunctive relief against such infringement pursuant to 35 U.S.C. § 283.

COUNT IV

26. GE Licensing repeats and realleges the allegations in paragraphs 1-13 as though fully set forth herein.

27. On information and belief, Agere has infringed, contributed to the infringement of, and/or induced infringement of, the '776 Patent by making, using, selling, offering for sale, or importing into the United States, or by aiding and abetting others to make, use, import into, offer for sale, or sell in the United States, products that incorporate or practice the invention of the '776 Patent.

28. On information and belief, Agere's infringement of the '776 Patent has been willful. Agere's continued infringement of the '776 Patent has damaged and will continue to damage GE Licensing.

29. On information and belief, unless enjoined by this Court, Agere will continue to infringe the '776 Patent, and GE Licensing will continue to suffer irreparable

harm for which there is no adequate remedy at law. Accordingly, GE Licensing is entitled to permanent injunctive relief against such infringement pursuant to 35 U.S.C. § 283.

PRAYER FOR RELIEF

WHEREFORE, GE Licensing respectfully requests that this Court enter judgment in its favor and grant the following relief:

A. Adjudge that Agere is infringing the GE Patents;

B. Adjudge that Agere's infringement of the GE Patents was willful, and that Agere's continued infringement of the GE Patents is willful;

C. Award GE Licensing damages in an amount adequate to compensate GE Licensing for Agere's infringement of the GE Patents, but in no event less than a reasonable royalty under 35 U.S.C. § 284;

D. Enter an order trebling any and all damages awarded to GE Licensing by reason of Agere's willful infringement of the GE Patents, pursuant to 35 U.S.C. § 284;

E. Enter an order awarding GE Licensing pre-judgment and post-judgment interest to the full extent allowed under the law, as well as its costs;

F. Enter an order finding that this is an exceptional case and award GE Licensing its reasonable attorneys' fees pursuant to 35 U.S.C. § 285;

G. Permanently enjoin Agere and its officers, agents, servants, employees and attorneys, subsidiaries, parent company, and all persons acting in active concert or participation with them, from further infringing, contributing to and/or inducing the infringement of the GE Patents, in accordance with 35 U.S.C. § 283, including:

(i) making, using, testing, selling, offering for sale, importing, leasing and licensing infringing products;

(ii) providing support and maintenance for infringing products; and

(iii) making new versions of, or enhancements to, infringing products; and

H. Award such other relief as the Court may deem appropriate and just under the circumstances.

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Dated: March 23, 2007
777761 / 31182

EXHIBIT A

U 7042968

THE UNITED STATES OF AMERICA

TO ALL TO WHOM THESE PRESENTS SHALL COME:

UNITED STATES DEPARTMENT OF COMMERCE
United States Patent and Trademark Office

December 08, 2006

THIS IS TO CERTIFY THAT ANNEXED HERETO IS A TRUE COPY FROM THE RECORDS OF THIS OFFICE OF:

U.S. PATENT: *5,048,054*
ISSUE DATE: *September 10, 1991*

By Authority of the
Under Secretary of Commerce for Intellectual Property
and Director of the United States Patent and Trademark Office

E. BORNETT
Certifying Officer

UNITED STATES PATENT AND TRADEMARK OFFICE
DEPARTMENT OF COMMERCE

United States Patent [19]

Eyuboglu et al.

[11] **Patent Number: 5,048,054**

[45] **Date of Patent: Sep. 10, 1991**

[54] **LINE PROBING MODEM**

[75] Inventors: **Vedat M. Eyuboglu**, Boston; **Ping Dong**, Norwood, both of Mass.

[73] Assignee: **Codex Corporation**, Mansfield, Mass.

[21] Appl. No.: **351,199**

[22] Filed: **May 12, 1989**

[51] **Int. Cl.**[5] **H04B 7/10**

[52] **U.S. Cl.** **375/8**; 375/10; 375/38; 375/100; 455/135; 455/62

[58] **Field of Search** 375/58, 8, 10, 38, 100; 455/62, 135-137; 370/79; 340/825.03

[56] **References Cited**

U.S. PATENT DOCUMENTS

4,309,773	1/1982	Johnson et al.	455/62
4,630,126	12/1986	Kaku et al.	358/280
4,679,227	7/1987	Hughes-Hartogs	379/98
4,756,007	7/1988	Qureshi et al.	375/37

OTHER PUBLICATIONS

Eyūboglu, "Detection of Severely Distorted Signals Using Decision Feedback Noise Prediction with Interleaving," IEEE Transaction of Communications, vol. 36, No. 4, Apr. 1988, pp. 401-409.

USSN 351,186, "Trellis Precoding for Modulation Systems".

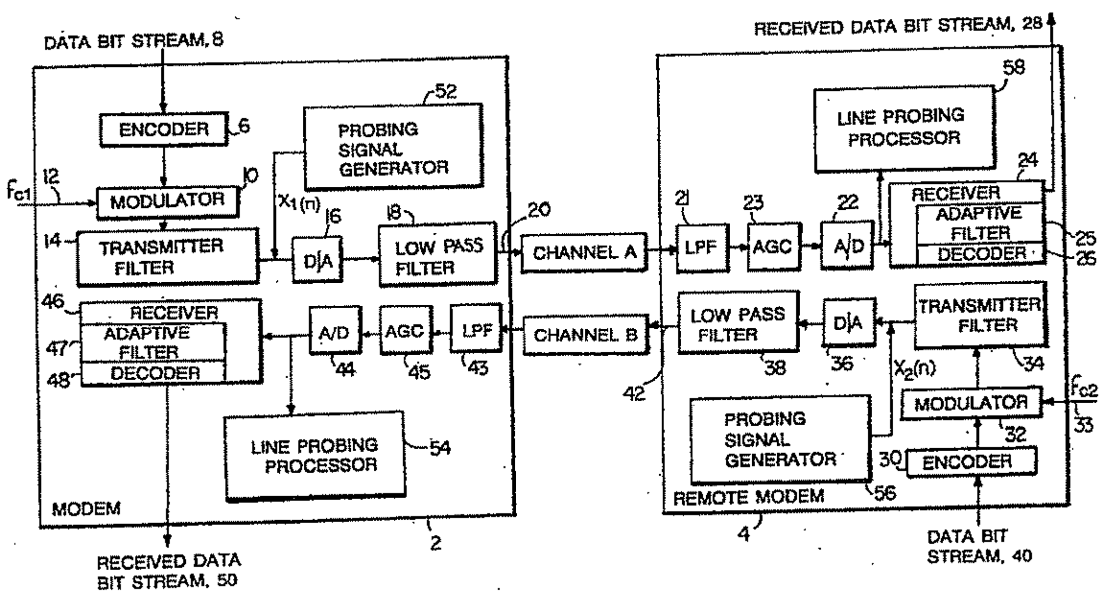
Kalet, "The Multitone Channel," IEEE Transactions on Communications, vol. 37, No. 2, Feb. 1989, pp. 119-124.

Primary Examiner—Benedict V. Safourek
Assistant Examiner—Tesfaldet Bocure
Attorney, Agent, or Firm—Fish & Richardson

[57] **ABSTRACT**

A modem for receiving data sent from a remote device over a communication channel by using a single carrier modulated signal, the modem including a receiver for receiving the modulated signal and for receiving a line probing signal sent by the remote device over the channel, the receiver being capable of receiving the modulated signal over any one of a plurality of frequency bands; a line probing processor for measuring characteristics of the channel based upon the received line probing signal; and a selector for selecting one of the plurality of frequency bands, said selection being based upon the measured characteristics of the channel, said selected frequency band to be used for receiving the modulated signal from the remote device.

71 Claims, 3 Drawing Sheets



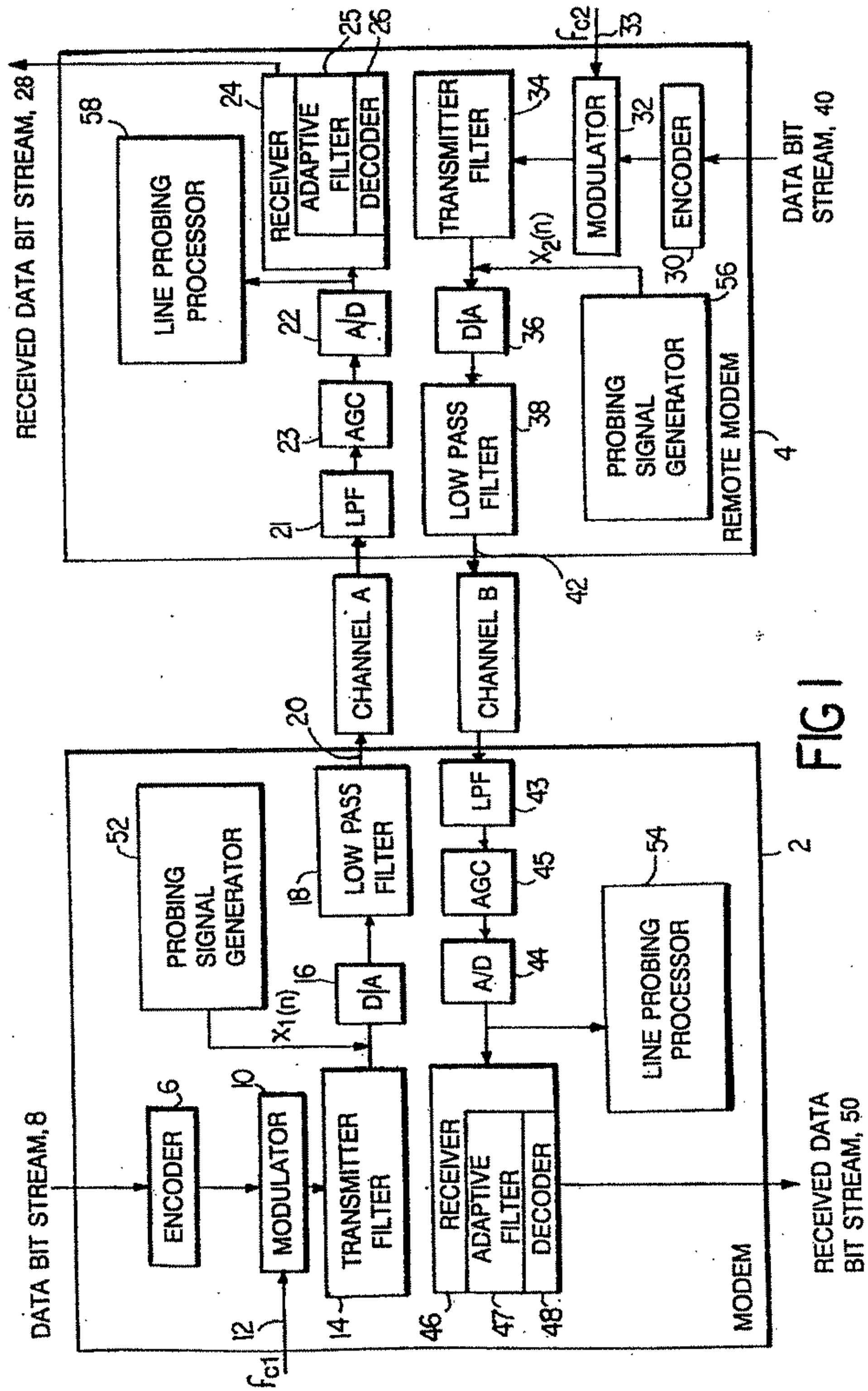

RECEIVED DATA BIT STREAM, 28
LINE PROBING PROCESSOR
58
RECEIVER
ADAPTIVE FILTER
DECODER
24
25
26
TRANSMITTER FILTER
34
MODULATOR
32
ENCODER
30
f_{c2}
33
DATA BIT STREAM, 40
A/D
22
AGC
23
LPF
21
D/A
36
$X_2(n)$
LOW PASS FILTER
38
PROBING SIGNAL GENERATOR
56
REMOTE MODEM
4
CHANNEL A
CHANNEL B
20
42
PROBING SIGNAL GENERATOR
52
LOW PASS FILTER
18
LPF
43
AGC
45
A/D
44
D/A
16
$X_1(n)$
LINE PROBING PROCESSOR
54
2
DATA BIT STREAM, 8
ENCODER
6
MODULATOR
10
TRANSMITTER FILTER
14
12
f_{c1}
RECEIVER
ADAPTIVE FILTER
DECODER
46
47
48
MODEM
RECEIVED DATA BIT STREAM, 50

FIG 1

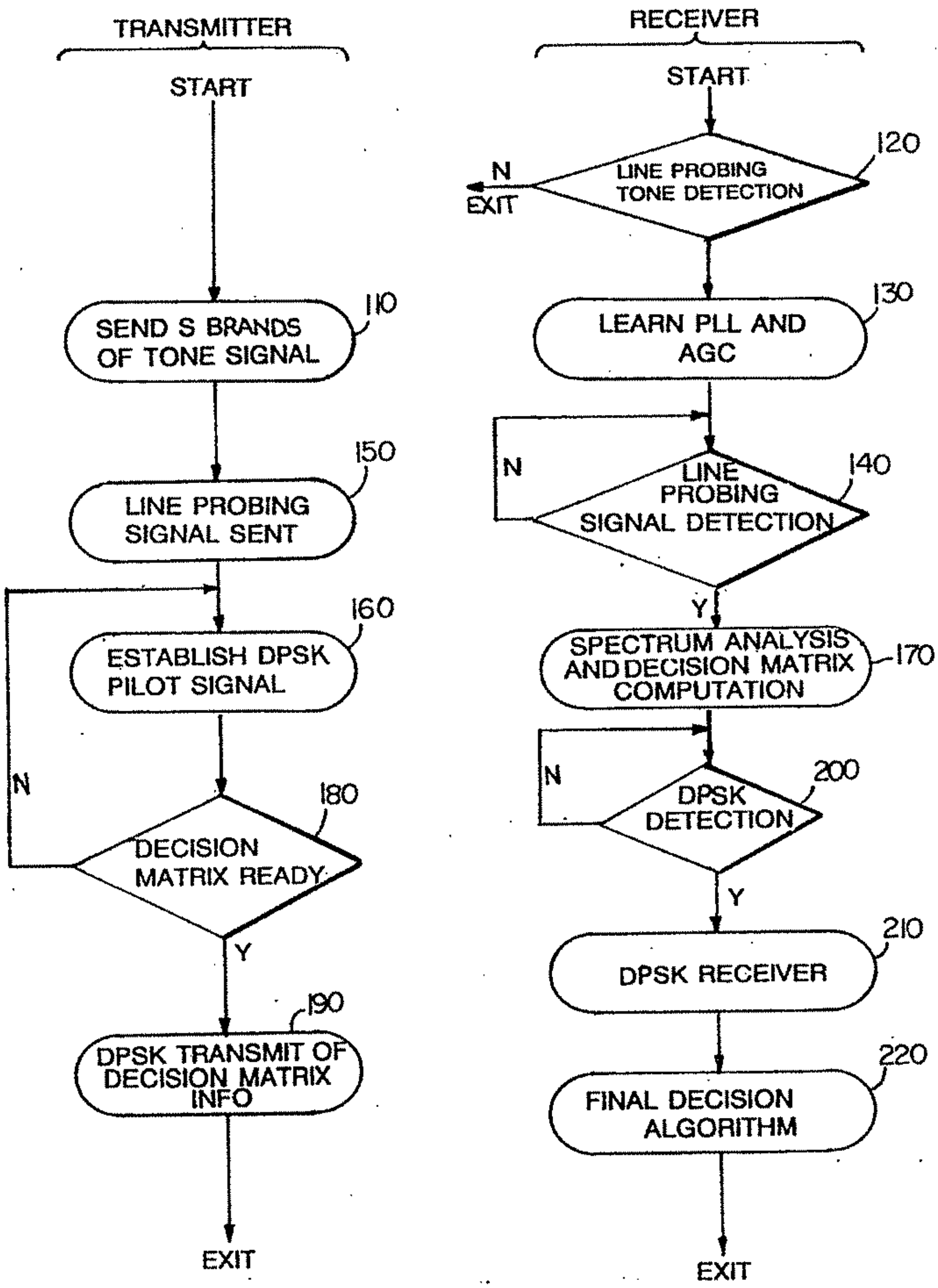

TRANSMITTER
START
SEND S BRANDS OF TONE SIGNAL
110
LINE PROBING SIGNAL SENT
150
ESTABLISH DPSK PILOT SIGNAL
160
N
DECISION MATRIX READY
180
Y
DPSK TRANSMIT OF DECISION MATRIX INFO
190
EXIT
RECEIVER
START
LINE PROBING TONE DETECTION
120
N
EXIT
LEARN PLL AND AGC
130
N
LINE PROBING SIGNAL DETECTION
140
Y
SPECTRUM ANALYSIS AND DECISION MATRIX COMPUTATION
170
N
DPSK DETECTION
200
Y
DPSK RECEIVER
210
FINAL DECISION ALGORITHM
220
EXIT

FIG. 2

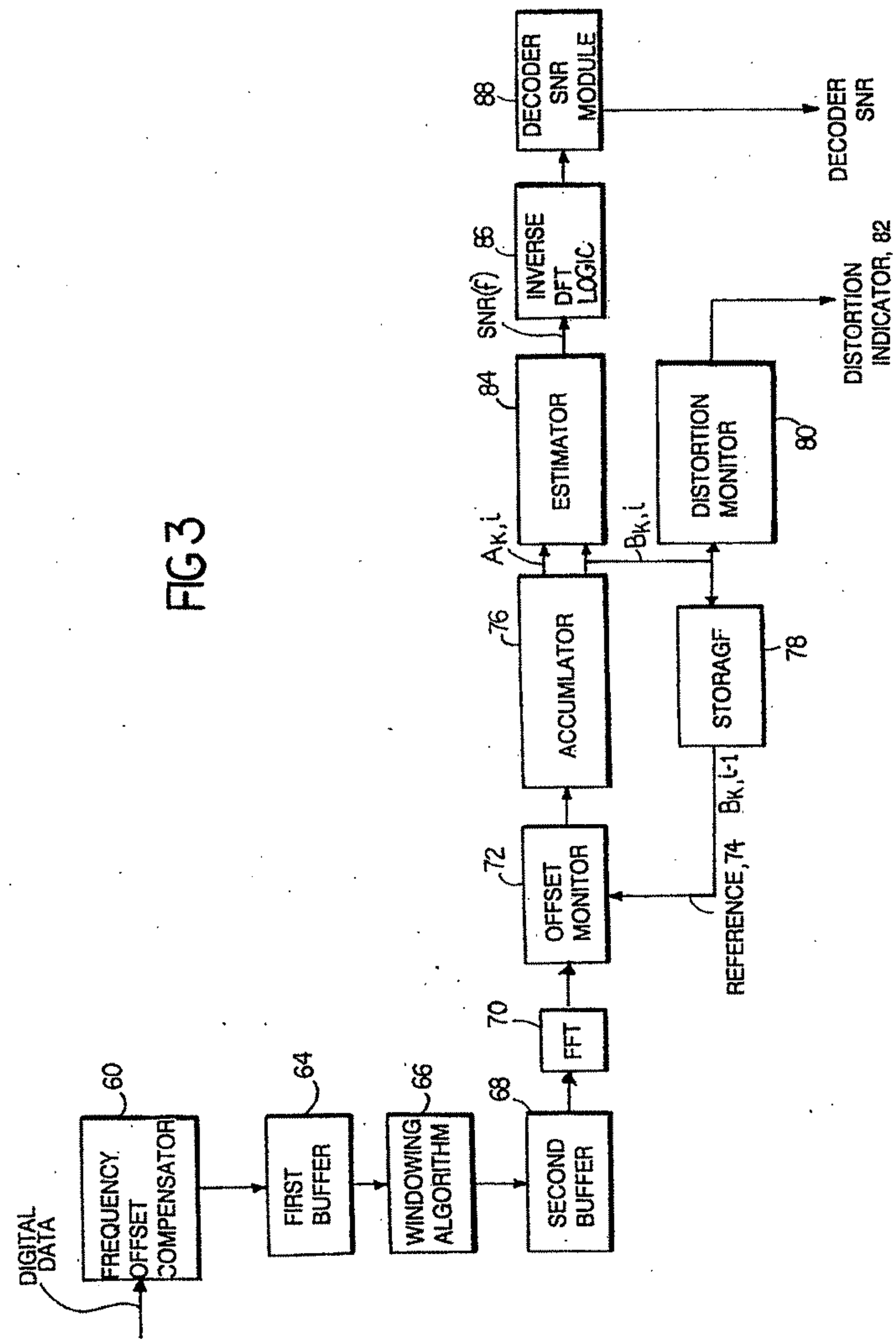

DIGITAL DATA
FREQUENCY OFFSET COMPENSATOR
60
FIRST BUFFER
64
WINDOWING ALGORITHM
66
SECOND BUFFER
68
FFT
70
OFFSET MONITOR
72
REFERENCE, 74
ACCUMLATOR
76
STORAGE
78
DISTORTION MONITOR
80
DISTORTION INDICATOR, 82
ESTIMATOR
84
INVERSE DFT LOGIC
86
DECODER SNR MODULE
88
DECODER SNR
SNR(f)

FIG 3

LINE PROBING MODEM

BACKGROUND OF THE INVENTION

This invention relates to data communication equipment or modems.

Modems are devices which employ digital modulation techniques to transmit binary data over analog band limited communication channels. High-speed modems commonly use linear modulation schemes such as quadrature amplitude modulation (QAM).

In linear modulation systems, binary information is collected in groups of M L bits (M is dimensionality and L is the bits/baud which may be fractional) and the resulting sequence is mapped into a sequence of complex-valued signal points, using some coding scheme. The complex sequence is filtered by a shaping filter to limit its bandwidth, and the real and imaginary components of the filtered signal points are used to amplitude modulate the quadrature components of a sinusoidal carrier of some frequency f_c. If the bit rate is R b/s, then $Q=R/L$ is the baud rate of the linear modulation system. The baud rate represents the minimum bandwidth required to transmit the modem signal without introducing distortion. (The actual bandwidth of the shaping filter may be larger, but it is typically proportional to the baud rate.) The baud rate and the carrier frequency together determine the transmission band.

The bandwidth efficiency of a linear modulation system is measured by L, the number of bits it transmits per baud. For fixed rate R, increasing L reduces the baud rate and thus the required bandwidth. However, increasing L also reduces the noise tolerance of the system. Therefore, for a given channel characteristic, there is an optimum tradeoff between the baud rate and the number of bits transmitted per baud.

On channels with a rectangular or brickwall-like spectrum and white noise, the baud rate must be chosen approximately equal to the channel bandwidth. On the other hand, if the channel spectrum shows gradual attenuation, it may be preferable to choose the baud rate large enough such that portions of the attenuated regions are included in the transmission band. A large baud rate results in increased distortion, however, an equalizer in the receiver can compensate for the distortion and the noise enhancement caused by equalization may be more than offset by the improved noise tolerance obtained with a smaller L.

In most commercial high-speed voiceband modems that are available today, the baud rate and carrier frequency and thus the transmission band is often fixed; e.g., $Q=2400$ Hz and $f_c=1800$ Hz. Recently, modems were introduced which offer multiple but manually selectable carrier frequencies. In either case, since channel characteristics show considerable variation between different lines or connections, with such modems it is difficult to achieve the best possible performance on all possible lines.

SUMMARY OF THE INVENTION

In general, in one aspect, the invention is a modem for receiving data sent from a remote device over a communication channel by using a single carrier modulated signal. The modem includes a receiver for receiving the modulated signal and for receiving a line probing signal sent by the remote device over the channel, the receiver being capable of receiving the modulated signal over any one of a plurality of frequency bands; a line probing processor for measuring characteristics of the channel based upon the received line probing signal; and a selector for selecting one of the plurality of frequency bands, said selection being based upon the measured characteristics of the channel, said selected frequency band to be used for receiving the modulated signal from the remote device.

In preferred embodiments, the measured characteristics include a frequency response of the channel and/or a signal-to-noise ratio of the channel measured at more than one frequency. The receiver includes an adaptive filter (which may implement trellis precoding) for providing a desired overall impulse response for the channel and at least some of the measured characteristics take into account the adaptive filter. And, the line probing signal is a substantially periodic signal.

Preferred embodiments also include the following features. The line probing processor includes a spectrum analyzer for generating discrete spectral representations of the received line probing signal; a module for estimating a frequency response for the channel based upon the discrete spectral representations of the received line probing signal, the frequency response being estimated at more than one frequency; and a module for estimating a power spectral density of channel noise based upon the discrete spectral representations of the received line probing signal. Also, the modulated signal is a linearly modulated signal (such as a quadrature amplitude modulated signal) and each one of said plurality of frequency bands is characterized by a corresponding baud rate and carrier frequency. The noise estimating module also estimates a power spectrum of the channel response based upon the discrete spectral representations of the received line probing signal and then computes a signal-to-noise ratio corresponding to the channel based upon both the power spectral density of channel noise and the power spectrum of the channel response. The noise estimating module performs weighted periodogram averaging to estimate the power spectral density of channel noise based upon the discrete spectral representations of the received line probing signal and also concurrently estimates the power spectrum of the channel response and the power spectral density of channel noise from the same received line probing signal. The modem also includes a transmitter for transmitting information based upon the measured characteristics to the remote device so that the remote device may identify one of said plurality of frequency bands based upon said transmitted information and then communicate said identified band to the receiver and wherein the selector selects the identified band as said selected band. If the received line probing signal may include an impairment (e.g. frequency offset and/or low frequency phase jitter), the line probing processor also includes an offset monitor for reducing effects of said impairment on the discrete representation of the received signal prior its being used to determine the power spectral density of channel noise. The offset monitor reduces the effects of said impairment by first estimating said impairment and by then rotating the discrete representation corresponding to a current period of the received line probing signal by an amount determined by the estimate of said impairment. The offset monitor estimates said impairment by comparing the discrete spectral representation corresponding to the current period of the received line probing signal to a reference signal derived from the discrete spectral

representations corresponding to at least one previous period of the received line probing signal. Also, the discrete spectral representations of the received line probing signal are M-point Discrete Fourier Transforms.

Preferred embodiments also include these additional features. The receiver is capable of receiving the modulated signal at any one of a plurality of bit rates and the modem further includes logic for selecting one of the plurality of different bit rates based upon the measured characteristics of the receiver channel, said selected bit rate to be used for receiving the modulated signal from the remote device. The line probing processor includes a spectrum analyzer for generating discrete spectral representations of the received line probing signal; and a module for computing a nonlinear distortion indicator based upon the discrete spectral representations of the received line probing signal. The receiver also includes a monitor circuit for measuring a power level of the received line probing signal and the measured characteristics includes a quantity derived from the received power level.

In general, in another aspect, the invention is a modem for transmitting data to a remote device over a communication channel by using a single carrier modulated signal. The modem includes a signal generator for generating a line probing signal; a transmitter for transmitting the modulated signal and for transmitting the line probing signal to the remote device over the channel, the transmitter being capable of transmitting the modulated signal over any one of a plurality of frequency bands; a receiver for receiving characteristics of the channel from the remote device, the characteristics being derived by the remote device from the transmitted line probing signal; and a selector for selecting one of the plurality of frequency bands, said selection being based upon the measured channel characteristics, the selected frequency band to be used for transmitting the modulated signal to the remote device.

In general, in yet another aspect, the invention is a modem for receiving data sent from a remote device over a communication channel by using a single carrier modulated signal. The modem includes a receiver for receiving the modulated signal and for receiving a line probing signal sent by the remote device over the channel, the receiver being capable of receiving the modulated signal at any one of a plurality of bit rates; a line probing processor for measuring characteristics of the channel based upon the received line probing signal; and a selector for selecting one of the plurality of bit rates, said selection being based upon the measured characteristics of the receiver channel, the selected bit rate to be used for receiving the modulated signal from the remote device.

The invention determines the best transmission band and maximum bit rate for the modem based upon an offline measurement of the characteristics of the particular channel to which the modem is connected. Thus, in comparison to other conventional modems which use a single carrier frequency modulation scheme, the invention makes better use of the available frequency band of the channel and does so from the beginning of data transmission. Moreover, for modems which utilize an adaptive rate system to establish and maintain optimum performance during the course of data communications, the invention provides an efficient way to initialize the adaptive rate system.

Furthermore, when using a QAM system, the invention achieves close to optimal utilization of the maximum theoretical capacity of the channel.

Other advantages and features will become apparent from the following description of the preferred embodiment, and from the claims.

DESCRIPTION OF THE PREFERRED EMBODIMENT

We first briefly describe the drawings.

FIG. 1 is a block diagram of a communication system which embodies the invention;

FIG. 2 is a flow chart depicting the operation of the line probing shown in FIG. 1; and

FIG. 3 is a block diagram of the portion of the modem which implements the spectrum analysis and the decision matrix computation step shown in FIG. 2.

STRUCTURE AND OPERATION

Referring to FIG. 1, a local modem 2, which is of a fourwire type, transmits information to a remote modem 4 over a channel A and receives information sent by the remote device 4 over a channel B. In local modem 2, an encoder 6 receives a data bit stream 8 and encodes the bits according to some coding scheme at a baud rate Q_1 selected from a set of available baud rates. Local modem 2 sends L_1 bits/baud where L_1 is selected based on a set of available bit rates. A modulator 10, using a carrier signal 12 of a frequency f_{c1} selected from a set of available carrier frequencies, modulates the output of encoder 6 and a transmit filter 14 produces pulse shaping to control the bandwidth of the transmit signal, all in accordance with some single carrier modulation scheme, e.g. quadrature amplitude modulation (QAM). Next, a digital-to-analog (D/A) converter 16 and a low pas filter 18 convert the digital transmit signal to an analog signal 20 which is transmitted over channel A to remote modem 4.

In remote modem 4, the received signal passes through a lowpass filter 21, an automatic gain control (AGC) circuit 23, an analog-to-digital (A/D) converter 22 and then a receiver 24, which includes an adaptive filter 25 followed by a decoder 26. Adaptive filter 25 provides a desired overall impulse response for decoder 26 which decodes the received signal according to the particular coding scheme used by local modem 2 to obtain an estimate of the transmitted data bit stream 8.

Remote modem 4 also includes an encoder 30, operating at a baud rate Q_2 selected from a set of available baud rates, a modulator 32 with a carrier signal 33 of frequency f_{c2}, selected from a set of available carrier frequencies, a transmitter filter 34, a digital-to-analog converter 36 and a low pass filter 38 which convert a data bit stream 40 into an analog signal 42 for transmission over channel B. Remote modem 4 sends L_2 bits per baud selected based on a set of available bit rates. Likewise, local modem 2 includes a low-pass filter 43, an automatic gain control circuit 45, an analog-to-digital converter 44 and a receiver 46. Similarly, receiver 46 includes an adaptive filter 47 which produces a desired overall impulse response for a decoder 48 which decodes the signal received over channel B to generate an estimate 50 of the data bit stream 8 transmitted by remote modem 4.

Local modem 2 includes a line probing signal generator 52, which generates a special probing signal sequence $x_1(n)$, and a line probing processor 54 which measures the quality of channel B. Likewise, remote

modem 4 includes a corresponding line probing signal generator 56, which generates a probing signal sequence $x_2(n)$, and a line probing processor 58 which measures the quality of channel A.

In general, local modem 2 sends its probing signal sequence $x_1(n)$ to line probing processor 58 of the remote modem 4, which uses the corresponding received signal sequence to compute the signal-to-noise ratio (SNR) for channel A as a function of frequency, i.e. $SNR_A(f)$. Then, for each combination of baud rate and carrier frequency available to it, remote modem 4 computes a corresponding decoder signal-to-noise ratio (which shall be defined shortly). For each baud rate, the carrier frequency which yields the best performance (i.e., as will be explained later, the highest decoder SNR) is saved along with a value representing the performance. These decisions are sent to local modem 2. Similarly, local modem 2 makes decisions based on a similar line measurement process and, in turn, sends its decisions to remote modem 4. Both modems then use the combined information to select the transmission bands (characterized by Q_1, Q_2, f_{c1}, and f_{c2}) and the transmission rates (determined by L_1 and L_2) to be used during subsequent data communications.

The probing sequences are periodic signals selected to fully and uniformly stimulate the entire channel over the spectrum of frequencies which may be useful for data communication. One such sequence consists of a group of equal amplitude tones which are evenly spaced within the frequency band of interest, namely, 100 to 3600 Hz. The frequency separation between the tones determines the frequency resolution of the resulting SNR measurements. It is desirable to select the phases of these tones so as to yield a relatively small peak-to-average ratio for the transmitted signal thereby reducing the possibility of driving the channel beyond its region of linear operation. The following is an example of one such probing sequence which satisfies these criteria:

$$x(n) = A \sum_{k_1}^{k_2} \cos(2\pi k f_\Delta n T_s + \theta_k),\ n = 0, 1, \ldots, P-1 \tag{1}$$

$$\theta_k = \pi(k-k_1)^2/(k_2-k_1) \tag{2}$$

where

A is a scaling constant;

n is a sampling interval index;

f_Δ is the frequency resolution;

k is a frequency interval index;

k_1 specifies the lowest frequency index included in the sequence;

k_2 specifies the highest frequency index included in the sequence;

T_s equals $1/f_s$, where f_s is the sampling rate; and

P equals f_s/f_Δ, the number of samples in one period of the line probing signal.

In the embodiment described herein, the sampling rate is 9600 Hz, P equals 256, the frequency resolution f_{66} is 37.5 Hz, k_1 equals 3 and k_2 equals 96 (i.e., covering a frequency range from 112.5 to 3600 Hz).

Line probing processors 54 and 58 employ the Fast Fourier Transform (FFT) technique to compute SNR(f) for their respective channels. They determine SNR(f) by measuring the frequency response, H(f), and the noise power spectral density, $\Phi(f)$, of the channel at the discrete frequencies excited by the probing signal, i.e. kf_Δ, where $k=k_1, \ldots, k_2$. Then, the processors 54 and 58 compute SNR(f) by using the following well-known relationship:

$$SNR(f) = |H(f)|^2/\Phi(f) \tag{3}$$

Before describing the steps of the measurement algorithm in detail, an explanation of the underlying rationale will be given.

In general, the real-valued received sequence sampled at times $(iP+n)T_s$ can be written in the form:

$$r(i, n) = x(n)*h(n) + w(i, n) \tag{4}$$

$$= y(n) + w(i, n)\ i = 0, 1, \ldots, N-1; \tag{5}$$

$$n = 0, 1, \ldots, P-1$$

where * signifies convolution, x(n) is a transmitted periodic probing signal, h(n) is the sampled channel impulse response, w(i,n) is a potentially colored noise sequence with a power spectral density of $\Phi(f)$, N is the number of observation periods, and i is an index for observation periods.

Since the probing signal x(n) has a flat spectrum within the frequency band of interest, the noiseless channel output y(n) has the following power spectrum:

$$|\Psi(kf_\Delta)|^2 = |H(kf_\Delta)|^2$$

where $H(kf_\Delta)$ is the Discrete Fourier Transform (DFT) of h(n) and $\Psi(kf_\Delta)$ is the DFT of y(n).

Estimates of both $H(kf_\Delta)$ and $\Phi(kf_\Delta)$ may be readily obtained from a P-point DFT of the received segment r(i,n) n=0,1, . . . , P-1, which is designated hereinafter as $R_i(kf_\Delta)$, i=0,1, . . . , N-1. If $\Psi(kf_\Delta)$ were precisely known, then an estimate of the noise spectrum $\Phi(kf_\Delta)$ could be obtained from the following periodogram average:

$$\Phi(kf_\Delta) \approx (1/N) \sum_{i=0}^{N-1} \left| R_i(kf_\Delta) - \Psi(kf_\Delta) \right|^2;\ k_1 \leqq k \leqq k_2. \tag{6}$$

It can be shown that the periodogram averaging yields an asymptotically unbiased and consistent estimate of the noise spectrum, $\Phi(kf_\Delta)$. That means, as the number of observation periods increases, the mean and variance of the error tend to zero.

Although $\Psi(kf_\Delta)$ is unknown, it may be estimated by using the following DFT averaging:

$$\Psi(kf_\Delta) \approx (1/N) \sum_{i=0}^{N-1} R_i(kf_\Delta);\ k_1 \leqq k \leqq k_2. \tag{7}$$

After substituting Eq. 7 into Eq. 6, with straightforward manipulations, Eq. 6 can be written in the following form:

$$N\Phi(kf_\Delta) \approx \Sigma_i |R_i(kf_\Delta)|^2 - |(1/\sqrt{N})\Sigma_i R_i(kf_\Delta)|^2 \tag{8}$$

With the substitutions

$$A_k = \Sigma_i |R_i(kf_{66})|^2;\ k_1 \leqq k \leqq k_2 \tag{9}$$

$$B_k = (1/\sqrt{N})\Sigma_i R_i(kf_{66});\ k_1 \leqq k \leqq k_2 \tag{10}$$

this further reduces to:

$$\Phi(kf_{66}) \approx (A_k - B_k^2)/N;\ k_1 \leq k \leq k_2. \quad (11)$$

Note that:

$$|H(kf_\Delta)|^2 \approx B_k^2;\ k_1 \leq k \leq k_2. \quad (12)$$

The line probing processors 54 and 58 use the above equations to simultaneously estimate the noise spectral density $\Psi(kf_\Delta)$ and the channel frequency response $H(kf_\Delta)$.

Every observation period, the algorithms accumulate and store $\Sigma_i|R_i(kf_\Delta)|^2$ and $(1/\sqrt{N})\Sigma_i R_i(kf_\Delta)$. After N observation periods, the results equal A_k and B_k, $k=k_1, \ldots, k_2$, respectively.

The estimate of $\Phi(kf_\Delta)$ at any given frequency kf_Δ may be affected by the spectral energy density at other frequencies and thus may be "biased". When computing the DFT's from the received signal r(i,n) with an FFT technique, the line probing processors 54 and 58 use windowing, a known spectrum analysis technique for improving the performance of simple periodogram averaging. In the embodiment described herein, a Hanning window of length 2P is used. The Hanning window has a raised cosine shape with a 100% roll-off. To reduce the calculation time associated with using a window of length 2P, two successive periods of received data are overlapped. Of course, a window of duration longer than 2P may be used to improve accuracy, however, this would increase the amount of computations required for completing the FFT calculations.

The received signal r(i,n) may also have a frequency offset that can substantially degrade the accuracy of the estimation of $R_i(kf_\Delta)$ and, in turn, the noise spectrum. Although, windowing reduces effects of the frequency offset, additional steps are taken to reduce it even further. In the noise-free situation, the presence of frequency offset causes the DFT for the current observation period to differ from the DFT for the preceding period by a constant phase factor. That is:

$$R_{i+1}(kf_\Delta)/R_i(kf_\Delta) = \text{phase factor} = \exp[j2\pi f_0 PT_s] \quad (13)$$

where f_0 is the frequency offset. This relationship is used to estimate the phase factor. Then, the estimated phase factor is used to rotate the DFT's to cancel the effects of the frequency offset.

In the embodiment described herein, line probing processors 54 and 58 use the accumulating estimates of B_k, $k=k_1, \ldots, k_2$, (which shall be designated as $B_{k,i}$, where i indicates the observation period) rather than the DFT from the previous period, to achieve basically the same results. That is, after the initial period of accumulation, the newly computed periodograms $R_i(kf_\Delta)$ are compared with the corresponding $B_{k,i-1}$ by taking the inner product:

$$E = \Sigma_k B_{k,i-1} R_i^*(kf_\Delta) \quad (14)$$

where * is the complex conjugate and the summation is over $k_1 \leq k \leq k_2$. The quantity E, a complex number, is normalized by using a polynomial approximation of the function $1/\sqrt{x}$ where x equals $[|Re[E]|^2 + |Im[E]|^2]$ and then it is used to rotate $R_i(kf_\Delta)$ before $R_i(kf_\Delta)$ is added to $B_{k,i-1}$ to produce $B_{k,i}$, a new estimate of B_k.

In addition to SNR(f), an important source of distortion for data transmission is non-linear distortion (NLD). NLD causes the energy in the transmitted frequency components to be spread over other frequencies. Although there are standard techniques for measuring NLD on telephone lines, it is desirable to obtain a rough estimate of NLD by using the line probing signals. Thus, to measure NLD, the line probing signal is slightly modified from the one described above. This is done by omitting some preselected frequency lines. Line probing processors 54 and 58 then measure the strength of the received line probing signals at these omitted frequencies and average those measurements to arrive at a rough estimate of NLD. As will be described later, the estimate of NLD is then taken into account in estimating the maximum achievable bit rates for the modems.

To account for the missing frequencies, line probing processors 54 and 58 estimate the missing values of $SNR(kf_\Delta)$ by averaging the values of $SNR(kf_\Delta)$ corresponding to frequency lines in the vicinity of the omitted frequency lines. The location of the omitted frequency lines are selected so that they are common to all transmission bands available to the modem and lie near the mid-range of such bands, where SNR(f) is likely to be relatively uniform. By selecting the omitted frequencies in this manner, the error caused by this approximation is kept small.

After N observation periods have elapsed and the estimates of A_k and B_k have been accumulated for $k_1 \leq k \leq k_2$, processors 54 and 58 compute $SNR(kf_\Delta)$ for the corresponding channels using Eqs. 11, 12 and 3, above. The computed $SNR(kf_\Delta)$ is then used to determine for each baud rate the carrier frequency which yields the best receiver performance. The way this is determined will now be described.

The decoder in each modem operates on a properly equalized signal, i.e. one which has passed through the receiver's adaptive filter. Thus, it is generally the SNR at the output of the adaptive filter, i.e. the decoder SNR, that is most relevant to the performance of the receiver. The decoder SNR is related to $SNR(kf_\Delta)$, and the precise relationship depends upon the type of adaptive filter used in the receiver.

In the described embodiment, modems 2 and 4 are equipped with a trellis precoding equalization system such as the scheme disclosed in the U.S. patent application entitled "Trellis Precoding for Modulation Systems, "by Eyuboglu and Forney, filed on an even date herewith and hereby incorporated by reference. In trellis precoding, each receiver 24 and 46 includes a fractionally-spaced minimum-mean-squared error linear equalizer whose output is sampled at the baud rate, followed by a linear prediction filter responsible for whitening the residual error sequence at the output of the linear equalizer. If $\{x_n'\}$ is the complex sequence transmitted by the trellis precoder, then the received sequence at the output of the prediction filter can approximately be written in the form

$$\{r_n\} = \{x_n'\} * \{h'_n\} + \{w_n\}, \quad (15)$$

where $\{h'_n\}$ is a causal (i.e., $h'_{nb} = 0$, $n>0$) overall impulse response and $\{w_n\}$ is a white error sequence of some variance σ^2. Here, it may be assumed that the filters are scaled such that $h_0=1$. Then, under the assumption that the error signal can be modeled as Gaussian and neglecting other small effects, it is known that the performance of the trellis precoder is given by

$$P_e \approx K\,Q(d_{min}/2\sigma) \quad (16)$$

where K is a constant, $Q(\alpha)$ is the Gaussian tail function given by

$$Q(\alpha) = (1/\sqrt{2\pi}) \int_{\infty}^{\alpha} \exp\{-\alpha^2/2\}\, d\alpha \tag{17}$$

and d_{min} is typically taken as the minimum distance between L allowable channel output sequences. The quantities K and d_{min} depend on the trellis code that is used in conjunction with the trellis precoder. Stated approximately, d_{min} decreases by a factor of $\sqrt{2}$ for every increment in L, the number of bits per baud, assuming that the average power of the precoder output is kept constant. The sampled decoder SNR may be defined as $d^2_{min}/2\sigma^2$. The baud rate or carrier frequency affect the decoder SNR only through the noise variance σ^2.

To determine the relationship between σ^2 and SNR(f), first note that at the equalizer output, after demodulation, the noise spectrum is given by

$$S_n(f) = \Phi(f)/|H(f)|^2,\ |f - f_c| < Q/2, \tag{18}$$

where Q equals the baud rate and it is assumed that the transmitted signal has zero excess-bandwidth. (Typically, highspeed modems use 10-12% excess bandwidth; however, experiments have shown that this has only a small effect on performance). Now, the autocorrelation sequence of the noise sequence can be computed as

$$g_n = T \int_{|f - f_c| < Q/2} \Phi(f)/|H(f)|^2\, e^{j2\pi fn/Q}\, df,\ n = 1, 2, \ldots \tag{19}$$

Since the spectra are measured at discrete frequencies kf_Δ, g_n can be approximated as follows:

$$g_n = a\, \Sigma \Phi(kf_\Delta)/|H(kf_\Delta)|^2 \exp(j2\pi kf_\Delta n/Q),\ n = 1, 2, \ldots \tag{20}$$

$$k_1(Q,f_c) < k < k_2(Q,f_c)$$

where a is some normalization factor and $k_1(Q,f_c)$ and $k_2(Q,f_c)$ are the frequency indices corresponding to the bandedges assuming the baud rate Q and carrier frequency f_c. Once $\{g_n\}$ are determined, σ^2 can be computed using well-known formulas for linear prediction. For example, if it can be assumed that the noise sequence is a first-order autoregressive (AR) process as described by Eyuboglu in "Detection of Severely Distorted Signals Using Decision Feedback Noise Prediction with Interleaving" IEEE Trans. on Communications, April, 1988, then, σ^2 is given by

$$\sigma^2 = g_0 - |g_1|^2/g_0. \tag{21}$$

Thus, by using Eqs. (20) and (21), line probing processors **54** and **58** can determine for each baud rate Q, the carrier frequency $f_c(Q)$ which yields the smallest noise power $\sigma^2(f_c(Q))$.

The computation of decoder SNR described above can be extended to higher order AR models using well known formulas for the minimum-mean-square error prediction error as described in Jayant and Noll, "Digital Coding of Waveforms" Prentice-Hall, 1984.

The description given above assumes that the fractionally-spaced linear equalizer has a sufficiently large span to reduce the effects of phase distortion to a negligible level. In highspeed modems that is often a reasonable assumption. In applications where this condition may not be satisfied, the effect of phase distortion has to be taken into account. Furthermore, if the assumption of a first-order AR model does not hold, then the residual noise sequence may be correlated and its effect on performance may have to be taken into consideration.

Having described the nature of the computations performed by processors **54** and **58**, the steps of the measurement algorithm, shown in FIG. **2**, will now be described in more detail. Local modem **2** starts the line probing process by transmitting a line probing tone signal at some fixed frequency for a fixed number of bauds (designated in FIG. **2** as S) over channel A (step **110**) while at the same time, remote modem **4** monitors channel A to detect the tone signal (step **120**).

After detecting the line probing tone, remote modem **4** initiates a period of tone training (step **130**). During this period, it uses a phase-locked loop (PLL) to learn the frequency offset in the incoming tone and at the same time it adjusts its AGC setting to achieve a desired signal level prior to A/D conversion. After a fixed amount of time, the receiver freezes both its AGC setting and its PLL and then switches to a transition detection state to detect the arrival of the wideband line probing signal transmitted from the local modem **2** (step **140**).

In the meantime, the transmitter in modem **2** continues to transmit its tone for at least S bauds and until it receives a reply tone from remote modem **4**. Then, processor **54** causes probing signal generator **52** to generate the above-describe special probing signal and transmits it to remote modem **4** for at least N periods (step **150**). After receiving N periods of the probing signal from remote modem **4**, modem **2** switches to a communication mode for sending line probing measurement results to remote modem **4** (step **160**).

Since the line probing algorithm implemented by remote modem **4** is the same as the line probing algorithm of modem **2**, the sequence of events in both modems **2** and **4** and their timing is also basically the same. Thus, remote modem **4** sends its tone signal, generates its probing signal for transmission to modem **2**, and enters a corresponding communication mode at about the same times as these events occur in modem **2**.

As soon as remote modem **4** detects the probing signal on channel A, processor **58** begins a spectrum analysis of the received probing signal (step **170**). First, line probing processor **58** measures the channel and noise spectra and from these computes SNR(f) and then the decoder SNRs.

The elements of the modem which perform the spectrum analysis and the decision matrix computation of step **170** are shown in FIG. **3**. After being converted to a digital signal, the received real-valued probing signal passes through a frequency offset compensator **60** which multiplies it by a complex-valued rotation factor which was derived from the frequency offset estimate obtained during the initial tone training described above. A first buffer **64** temporarily stores the rotated digital signal for subsequent processing.

After buffer **64** has received data corresponding to two periods of the line probing signal, a windowing algorithm **66** applies a Hanning window to the two periods of rotated data stored in buffer **64** to produce a frame of windowed data which consists of 2P complex-valued samples. These are then stored in a second buffer

68. Next, FFT algorithm 70 computes a P-point DFT from the stored frame of windowed data.

After each new period of data is received and stored in first buffer 64, windowing algorithm 66 uses the stored data, along with the data from the preceding period, to compute a new frame of windowed data. FFT algorithm 70 then computes a new P-point DFT using the new frame of windowed data. In other words, for each period of the received probing signal, a new P-point DFT is generated from the two most recent periods of data. Thus, one period of data is used to compute the DFT for two successive periods.

The P-point DFT's from FFT algorithm 70 are passed to a frequency offset monitor 72, which first estimates and then reduces any uncancelled frequency offset which may be present. Offset monitor 72 estimates the amount of uncancelled frequency offset by comparing each computed DFT against a reference 74 which corresponds to accumulated DFT's from previous observation periods. Offset monitor 72 then rotates the elements of the DFT by an amount that corresponds to the estimated uncancelled frequency offset for that DFT, thereby generating the rotated DFT, $R_i'(kf_\Delta)$.

Next, an accumulator 76 receives these rotated DFT's and uses them to generate the quantities A_k and B_k-as follows. Initially, accumulator 76 sets both $A_{k,0}$ and $B_{k,0}$ to zero. During the i^{th} period, when $R_i'(kf_\Delta)$ has been compiled, for each k in the range $k_1 \leqq k \leqq k_2$, accumulator 76 computes the squared magnitude of $R_i'(kf_\Delta)$ and adds the result to the stored value for $A_{k,i-1}$ to generate $A_{k,i}$. Accumulator 76 also divides the $R_i'(kf_\Delta)$ by $\sqrt{N}$ and adds this result to the stored value for $B_{k,i-1}$ to generate $B_{k,i}$. $A_{k,i}$ and $B_{k,i}$ are accumulated in this manner for N observation periods. (Note that $A_{k,i}$'s are real numbers, whereas $B_{k,i}$'s are complex numbers.)

The $B_{k,i}$'s generated during each observation period are stored in storage element 78 for use as a reference signal 76 during the next observation period. That is, for the i^{th} observation period, reference signal 76 is equal to $B_{k,i-1}$ computed during the previous observation period. Offset monitor 72 uses reference signal 76 to compute Eq. 14 described earlier. That is, offset monitor 72 calculates the complex conjugate of the current P-point DFT from FFT algorithm 70, i.e. $R_i^*(kf_\Delta)$, and then computes the inner product of $B_{k,i-1}$ and $R_i^*(kf_\Delta)$. The inner product is then normalized to arrive at an estimate of the phase offset. It should be noted that by using this approach, the modem can compensate for any residual frequency offset as well as track small amounts of low frequency phase jitter.

Since preselected frequencies were omitted from the line probing signal, the values of B_k at the locations of the omitted frequencies provide a measure of the non-linear distortion associated with the channel. Thus, after $B_{k,N}$ has been determined for N observation periods, a distortion monitor 80 estimates the non-linear distortion by squaring the amplitudes of $B_{k,N}$'s corresponding to omitted frequencies and computing the average of the resulting squared amplitudes. The average is then supplied by monitor 80 as a non-linear distortion indicator 82.

Using the $A_{k,N}$'s and $B_{k,N}$'s from accumulator 76, an estimator 84 then estimates the noise spectrum, $\Phi(kf_\Delta)$, and the channel spectrum, $|H(kf_\Delta)|^2$, for all frequencies used in the probing signal. Estimator 84 accomplishes this by first computing $|B_{k,N}|^2$ and then using Eqs. 11 and 12, described above. Using the estimates for the noise spectrum and the channel spectrum, estimator 84 then computes $SNR(kf_\Delta)$ in accordance with Eq. 3, described above. For the omitted frequency lines, estimator 84 approximates their SNR value by averaging the values of $SNR(kf_\Delta)$ over frequency lines in their vicinity.

Transform logic 86 receives the resulting values for $SNR(kf_\Delta)$ from estimator 84 and computes the inverse DFT specified by Eq. 20, above. The output of transform logic 84 is the noise autocorrelation function described earlier. Finally, a decoder SNR module 88 calculates the decoder SNR from the output of transform logic 86 according to Eq. 21.

Using the approach just described and also as part of step 170 shown in FIG. 2, remote modem 4 then makes a number of local decisions. Such local decisions help reduce the amount of information that needs to be exchanged with local modem 2. (Note that the local decision procedures to be described are the same for both modems 2 and 4. In particular and in accordance with the approach described earlier, remote modem 4 uses $SNR(kf_\Delta)$ to compute the decoder SNR's for each baud-rate/carrier-frequency combination available to it and then selects the best carrier frequency for each of the available baud rates.

The computed decoder SNR's, the non-linear distortion indicator, the signal power level of the received signal, as reflected by the AGC setting, and a user specified error performance requirement are then used to determine for each available baud-rate Q (using the best-carrier-frequency) and the maximum number of bits per baud $L_1(Q_1)$ that remote modem 4 can receive at without violating the performance requirement. To determine $L_1(Q_1)$, modem 4 uses a precomputed conversion table which is indexed on the basis of the above-identified information.

Basically, the conversion table depends upon the modem's modulation scheme, the coding gain of the coding scheme used, the way those schemes are implemented, and the error performance requirements. If trellis precoding is employed, the relationship between performance and decoder SNR is approximately described by Eq. 16 above. Nonlinear distortion and receive power level, however, modify that relationship somewhat. The actual entries in the conversion table can be derived, in part, from empirical observations and experiments in which the relationship between performance and the decoder SNR, NLD, and receive power level is measured for the particular type of modem being used.

After $L_1(Q_1)$ is obtained for all baud rates, remote modem 4 can calculate, for each of the available baud rates Q_1, the maximum bit rate $R_1(Q_1)$ it can receive from local modem 2 according to the following relation:

$$R_1(Q_1) = Q_1 \times L_1(Q_1)$$

When the spectrum analysis is complete, line probing processor 58 stores the results in a decision matrix. Upon completing the entries to the decision matrix, processor 58 indicates that its matrix is ready (step 180) and remote modem 4 transmits the information contained in its matrix to local modem 2 over channel B (step 190).

In each modem, the user or the network system may specify a maximum receive bit rate, R_{max}, and a minimum receive bit rate, R_{min}. This user-specified operating range is taken into account when the modem deter-

mines the decision matrix entries. Thus, if the selected bit rate for a particular baud rate is greater than R_{max}, then modem 2 sets it to R_{max}. Whereas, if the selected bit rate for a particular baud rate is less than R_{min}, then modem 2 sets it to R_{min} and also sets a flag associated with that baud rate to indicate that the performance requirement cannot be met at that baud rate. Note that a user can force a desired bit rate by setting $R_{max}=R_{min}$=desired rate.

The user also has the option to disable some (but not all) of the available baud rates. For example, the user may wish to operate at a specific baud rate. A second flag corresponding to each of the available baud rates is set to indicate whether that baud rate is disabled.

Of course, other constraints, besides those mentioned above, may also limit the communication options available to modems 2 and 4. For example, a user may require symmetric baud rates or symmetric bit rates in both transmission directions. Such additional constraints are stored in the decision matrix of the corresponding modem and are sent to the other modem along with other relevant information.

Specifically, during step 190, the following information is sent from the remote modem 4 to local modem 2:

a) the maximum bit rate at which the remote modem can receive for each of the available baud rates;

b) the best carrier frequency to be used for each of the available baud rates;

c) a flag for each of the available baud rates indicating whether the performance requirement can be met;

d) a flag for each of the available baud rates, indicating whether that baud rate is disabled in the remote modem;

e) a flag to indicate whether symmetric bit rates are required for both directions of transmission; and

f) a flag to indicate whether symmetric baud rates are required for both directions of transmission.

Naturally, some synchronization bits to indicate to beginning of the data and parity bits for error checking may also be transmitted during this information exchange phase

Since the decision matrix information is short, it can be transmitted quickly and reliably by using a simple, low-speed, robust modulation scheme which does not require a long training procedure. In this embodiment, this is achieved by using Differential-Phase-Shift-Keying (DPSK) at 300 bps. Other reliable modulation schemes such as low-speed Frequency-Shift-Keying (FSK) may also be employed.

Processor 54 monitors channel B for the presence of a DPSK signal carrying the decision matrix for channel A from remote modem 4 (step 200). When the DPSK signal is detected on channel B, processor 54 activates a DPSK receiver in modem 2 that includes a timing recovery circuit to provide correct sampling phase and then processor 54 decodes the decision matrix from the DPSK signal (step 210).

The DPSK receiver first looks for a synchronization pattern from the received bit stream. Once the pattern is detected, the receiver decodes the subsequent bits carrying the decision matrix. At the same time, the receiver also computes a parity check. At the end of DPSK transmission, this parity is compared with the one received from the remote modem. If they do not agree, a DPSK transmission error is flagged.

After modems 2 and 4 have exchanged their decision matrix information in this manner, they have complete information about channels A and B, including the operational constraints. Modem 2 then executes a final decision algorithm to select the carrier frequencies baud rates and bit rates to be used for communication over channels A and B (step 220). Since both modems 2 and 4 have the same information, they make the same selections and a further exchange of final decisions is not required.

The final decision algorithm first checks if any one of the modems required symmetric bit rates, or symmetric baud rates. If one of modems requested a symmetric baud rate or a symmetric bit rate, the request is enforced on both modems.

More specifically, if symmetric baud rates are required, the decision algorithm checks whether there are allowable baud rates common to both modems (i.e., baud rates which both satisfy the performance requirements and are allowed). If such baud rates exist, modems select the baud rate that maximizes the smaller of the two bit rates. When there are no allowable baud rates common to both modems, the decision algorithm includes baud rates that do not satisfy the performance requirement to find a common baud rate. A possible criterion for determining the reasonable baud rate may be to use a baud rate whose carrier frequency is closest to the center of the frequency band. Since both modems use the same criterion to choose this baud rate, they should both reach the same conclusion and no confusion will occur.

On the other hand, if symmetric baud rates are not required, the decision algorithm chooses from all of local modem's allowed baud rates the receiver baud rate that maximizes local modem's receiver bit rate and it chooses from all of remote modem's allowed baud rates the transmitter baud rate that maximizes remote modem's receiver bit rate.

After the transmitter and the receiver baud rates for the two modems are finalized, the best carrier frequencies associated with these baud rates are used as the transmitter and receiver carrier frequencies. Unless a symmetric bit rate is required, the maximum bit rate for each of those baud rates is used as the transmission bit rate for the corresponding modem. If a symmetric bit rate is required, the lower of the two bit rates (i.e. the bit rate for the local modem receiver and the bit rate for the remote modem receiver) is used as the common bit rate.

The main outputs of the line probing processor are the transmitter and receiver baud rates, Q_1 and Q_2, the transmitter and receiver carrier frequencies, f_{c1} and f_{c2}, the transmitter and receiver bit rates, R_1 and R_2, as well as an error code, which may indicate some unexpected error during the line probing process (such as failure in detecting the line probing signal, failure in synchronization, DPSK transmission error, etc.).

After the line probing is completed, modems go through a training at the selected baud rate and carrier frequencies and subsequently begin exchanging actual data at the selected rates.

Although the described embodiment used a four-wire type modem, it should be understood that this invention could also be carried out using a two-wire type modem. During full-duplex communication using a two-wire type modem, the received signals may, of course, include echo. For purposes of conducting the line probing measurements, it is desirable to avoid echoes in the received signal and this can easily be accomplished by having the modems conduct the line probing measurements sequentially rather than concurrently, as in the above-described embodiment.

Other embodiments may include the following features. The selection of the number of bits per baud may be based on the measurement of impairments in addition to or other than NLD and receive level. Also, in certain applications, no operational constraints may be necessary, in which case, the information exchange between the local modem and the remote device may be simplified. For example, each modem could immediately select its respective bit rate and the transmission band based upon its channel measurement and then exchange its final decision with the other modem. In addition, other exchange protocols may be used. Further, the selection of the transmission band may be based only on the measured frequency response of the channel and may not require measurement of the noise spectrum. Also, baseband data transmission may be employed instead of the passband transmission used in the described embodiment.

Other embodiments are within the following claims.

What is claimed is:

1. A modem for receiving data sent from a remote device over a communication channel by using a single carrier modulated signal, the modem comprising:

a. a receiver for receiving the modulated signal and for receiving a line probing signal sent by the remote device over the channel, the receiver being capable of receiving the modulated signal over any one of a plurality of frequency bands, said line probing signal simultaneously stimulating more than one of said plurality of frequency bands;

b. a line probing processor for measuring characteristics of the channel based upon the received line probing signal; and

c. a selector for selecting one of the plurality of frequency bands, said selection being based upon the measured characteristics of the channel, said selected frequency band to be used for receiving the modulated signal from the remote device.

2. The modem of claim **1** wherein the line probing processor comprises:

a. a spectrum analyzer for generating discrete spectral representations of the received line probing signal; and

b. a module for estimating a frequency response for the channel based upon the discrete spectral representations of the received line probing signal, the frequency response being estimated at more than one frequency.

3. The modem of claim **1** wherein the line probing processor comprises:

a. a spectrum analyzer for generating discrete spectral representations of the received line probing signal; and

b. a module for estimating a power spectral density of channel noise based upon the discrete spectral representations of the received line probing signal.

4. The modem of claims **1**, **2**, or **3** wherein the modulated signal is a linearly modulated signal and wherein each one of said plurality of frequency bands is characterized by a corresponding baud rate and carrier frequency, the modulated signal from the remote device being received at the corresponding baud rate associated with said selected frequency band.

5. The modem of claim **4** wherein the carrier frequency of two or more of said plurality of frequency bands are the same.

6. The modem of claim **4** wherein the linearly modulated signal is a quadrature amplitude modulated signal.

7. The modem of claim **1** wherein the measured characteristics include a frequency response of the channel.

8. The modem of claim **1** wherein the receiver includes an adaptive filter for providing a desired overall impulse response to a decoder and wherein at least some of the measured characteristics take into account the adaptive filter.

9. The modem of claim **8** wherein the adaptive filter is used in conjunction with trellis precoding.

10. The modem of claim **3** wherein the module performs weighted periodogram averaging to estimate the power spectral density of channel noise based upon the discrete spectral representations of the received line probing signal.

11. The modem of claim **1** wherein the line probing signal is a substantially periodic signal.

12. A modem for receiving data sent from a remote device over a communication channel by using a single carrier modulated signal, the modem comprising:

a. a receiver for receiving the modulated signal and for receiving a line probing signal sent by the remote device over the channel, the receiver being capable of receiving the modulated signal over any one of a plurality of frequency bands, each one of said plurality of frequency bands being characterized by a corresponding baud rate and carrier frequency;

b. a line probe processor for measuring characteristics of the channel based upon the received line probing signal; and

c. a selector for selecting one of the plurality of frequency bands, said selection being based upon the measured characteristics of the channel, the modulated signal from the remote device being received at the corresponding baud rate associated with said selected frequency band.

13. The modem of claim **12** wherein the line probing processor comprises:

a. a spectrum analyzer for generating discrete spectral representations of the received line probing signal; and

b. a module for estimating a frequency response for the channel based upon the discrete spectral representations of the received line probing signal, the frequency response being estimated at more than one frequency.

14. The modem of claim **12** wherein the line probing processor comprises:

a. a spectrum analyzer for generating discrete spectral representations of the received line probing signal; and

b. a module for estimating a power spectral density of channel noise based upon the discrete spectral representations of the received line probing signal.

15. The modem of claim **14** wherein the module performs weighted periodogram averaging to estimate the power spectral density of channel noise based upon the discrete spectral representations of the received line probing signal.

16. The modem of claim **15** wherein the module also estimates a power spectrum of the channel response based upon the discrete spectral representations of the received line probing signal and then computes a signal-to-noise ratio corresponding to the channel based upon both the power spectral density of channel noise and the power spectrum of the channel response.

17. The modem of claim **16** wherein the power spectrum of the channel response and the power spectral

density of channel noise are estimated concurrently from the same received line probing signal.

18. The modem of claim 16 wherein the receiver further comprises an adaptive filter for providing a desired overall impulse response to a decoder and the signal-to-noise ratio is determined relative to the output of the adaptive filter.

19. The modem of claim 18 wherein the adaptive filter is used in conjuction with trellis precoding.

20. The modem of claim 1 further comprising a transmitter for transmitting information based upon the measured characteristics to the remote device so that the remote device may identify one of said plurality of frequency bands based upon said transmitted information and then communicate said identified band to the receiver and wherein the selector selects the identified band as said selected band.

21. The modem of claim 12 wherein the measured characteristics include a frequency response of the channel.

22. The modem of claim 12 wherein the receiver includes an adaptive filter for providing a desired overall impulse response for the channel and wherein at least some of the measured characteristic take into account the adaptive filter.

23. The modem of claim 12 wherein the measured characteristics further include a signal-to-noise ratio of the channel measured at more than one frequency.

24. The modem of claim 12 wherein the line probing signal is a substantially periodic signal.

25. The modem of claim 11 or 24 wherein the line probing signal is of the form:

$$x(t)=A\Sigma_k\cos(2\pi kf_\Delta t+\theta_k),$$

where

$x(t)$ represents the line probing signal;

t is a time variable;

A is a constant;

f_Δ is a frequency resolution;

θ_k is a phase angle;

k is a frequency interval index which belongs to a subset of the integers ranging from k_1 through k_2;

k_1 specifies the lowest frequency index included in the line probing signal; and

k_2 specifies the highest frequency index included in the line probing signal.

26. The modem of claim 25 wherein the phase angles θ_k of the line probing signal are selected to achieve a small peak-to-average ratio of the line probing signal.

27. The modem of claim 25 wherein the phase angles θ_k of the line probing signal are equal to:

$$\theta_k=\pi(k-k_2)^2(k-k_1)$$

28. The modem of claim 12 wherein the modulated signal from the remote device is received at the corresponding baud rate and carrier frequency associated with said selected frequency band, and wherein at least some of the carrier frequencies associated with said plurality of frequency bands are different.

29. The modem of claim 1 wherein the measured characteristics further include a signal-to-noise ratio of the channel measured at more than one frequency.

30. The modem of claim 3 wherein received line probing signal may include an impairment (e.g. frequency offset and/or low frequency phase jitter) and wherein the line probing processor further comprises an offset monitor for reducing effects of said impairment on the discrete representation of the received signal prior its being used to determine the power spectral density of channel noise.

31. The modem of claim 30 wherein the line probing signal is substantially periodic and the discrete spectral representations are generated for each period of the received line probing signal and wherein the offset monitor reduces the effects of said impairment by first estimating said impairment and by then rotating the discrete representation corresponding to a current period of the received line probing signal by an amount determined by the estimate of said impairment.

32. The modem of claim 31 wherein the offset monitor estimates said impairment by comparing the discrete spectral representation corresponding to the current period of the received line probing signal to a reference signal derived from the discrete spectral representations corresponding to at least one previous period of the received line probing signal.

33. The modem of claim 2 or 3 wherein the discrete spectral representations of the received line probing signal are M-point Discrete Fourier Transforms.

34. The modem of claim 1 wherein the receiver is also capable of receiving the modulated signal at any one of a plurality of bit rates and wherein the modem further comprises logic for selecting one of the plurality of different bit rates based upon the measured characteristics of the receiver channel, said selected bit rate to be used for receiving the modulated signal from the remote device.

35. The modem of claim 34 wherein the line probing processor comprises:

a. a spectrum analyzer for generating discrete spectral representations of the received line probing signal; and

b. a module for computing a nonlinear distortion indicator based upon the discrete spectral representations of the received line probing signal.

36. The modem of claim 34 wherein the receiver comprises a monitor circuit for measuring a power level of the received line probing signal and wherein the measured characteristic includes a quantity derived from the received power level.

37. A modem for transmitting data to a remote device over a communication channel by using a single carrier modulated signal, the modem comprising:

a. a signal generator for generating a line probing signal;

b. a transmitter for transmitting the modulated signal and for transmitting the line probing signal to the remote device over the channel, the transmitter being capable of transmitting the modulated signal over any one of a plurality of frequency bands, said line probing signal simultaneously stimulating more than one of said plurality of frequency bands;

c. a receiver for receiving characteristics of the channel from the remote device, the characteristics being derived by the remote device from the transmitted line probing signal; and

d. a selector for selecting one of the plurality of frequency bands, said selection being based upon the measured channel characteristics, the selected frequency band to be used for transmitting the modulated signal to the remote device.

38. The modem of claim 37 wherein the modulated signal is a linearly modulated signal and wherein each

one of said plurality of frequency bands is characterized by a corresponding baud rate and carrier frequency.

39. The modem of claim 38 wherein the linearly modulated signal is a quadrature amplitude modulated signal.

40. The modem of claim 37 wherein the measured characteristics include a frequency response of the channel.

41. The modem of claim 37 wherein the measured characteristics further include a signal-to-noise ratio of the channel measured at more than one frequency.

42. The modem of claim 37 wherein the line probing signal is a substantially periodic signal.

43. The modem of claim 42 wherein the line probing signal is of the form:

$$x(t)=A\Sigma_k\cos(2\pi kf_\Delta t+\theta_k),$$

where

x(t) represents the line probing signal;

t is a time variable;

A is a constant;

f_Δ is a frequency resolution;

θ_k is a phase angle;

k is a frequency interval index which belongs to a subset of the integers ranging from k_1 through k_2;

k_1 specifies the lowest frequency index included in the line probing signal; and

k_2 specifies the highest frequency index included in the line probing signal.

44. The modem of claim 43 wherein the phase angles θ_k of the line probing signal are selected to achieve small peak-to-average ratio for the line probing signal.

45. The modem of claim 43 wherein the phase angles θ_k of the line probing signal are equal to:

$$\theta_k=\pi(k-k_2)^2/(k_2-k_1)$$

46. A modem for receiving data sent from a remote device over a communication channel by using a single carrier modulated signal, the modem comprising:

a. a receiver for receiving the modulated signal and for receiving a line probing signal sent by the remote device over the channel, the receiver being capable of receiving the modulated signal at any one of a plurality of bit rates;

b. a line probing processor for measuring characteristics of the channel based upon the received line probing signal; and

c. a selector for selecting one of the plurality of bit rates, said selection being based upon the measured characteristics of the receiver channel, the selected bit rate to be used for receiving the modulated signal from the remote device.

47. The modem of claim 46 wherein the line probing processor comprises:

a. a spectrum analyzer for generating discrete spectral representations of the received line probing signal; and

b. a module for estimating a frequency response for the channel based upon the discrete spectral representations of the received line probing signal, the frequency response being estimated at more than one frequency.

48. The modem of claims 46 wherein the line probing processor comprises:

a. a spectrum analyzer for generating discrete spectral representations of the received line probing signal; and

b. a module for estimating a power spectral density of channel noise based upon the discrete spectral representations of the received line probing signal.

49. The modem of claims 46, 47, or 48 wherein the modulated signal is a linearly modulated signal.

50. The modem of claim 49 wherein the measured characteristics include a frequency response of the channel.

51. The modem of claim 49 wherein the linearly modulated signal is a quadrature amplitude modulated signal.

52. The modem of claim 48 wherein the module performs weighted periodogram averaging to estimate the power spectral density of channel noise from the discrete spectral representations of the received line probing signal.

53. The modem of claim 52 wherein the module also estimates a power spectrum of the channel response based upon the discrete spectral representations of the received line probing signal and then computes a signal-to-noise ratio corresponding to the channel based upon both the power spectral density of channel noise and the power spectrum of the channel response.

54. The modem of claim 53 wherein the power spectrum and the spectral power density of channel noise are estimated concurrently from the same received line probing signal.

55. The modem of claim 48 wherein received line probing signal may include an impairment (e.g. frequency offset and/or low frequency phase jitter) and wherein the line probing processor further comprises an offset monitor for reducing effects of said impairment on the discrete representation of the received signal prior its being used to determine the power spectral density of channel noise.

56. The modem of claim 55 wherein the line probing signal is substantially periodic and a discrete spectral representation is generated for each period of the received line probing signal and wherein the offset monitor reduces the effects of said impairment by first estimating said impairment and by then rotating the discrete representation corresponding to a current period of the received line probing signal by an amount determined by the estimate of said impairment.

57. The modem of claim 56 wherein the offset monitor estimates said impairment by comparing the discrete spectral representation corresponding to the current period of the received line probing signal to a reference signal derived from the discrete spectral representations corresponding to at least one previous period of the received line probing signal.

58. The modem of claim 46 wherein the measured characteristics include a signal-to-noise ratio of the channel measured at more than one frequency.

59. The modem of claim 46 wherein the line probing signal is a substantially periodic signal.

60. The modem of claim 59 wherein the line probing signal is of the form:

$$x(t)=A\Sigma_k\cos(2\pi kf_\Delta t+\theta_k),$$

where

x(t) represents the line probing signal;

t is a time variable;

A is a constant;

f_Δ is a frequency resolution;

θ_k is a phase angle;

k is a frequency interval index which belongs to a subset of the integers ranging from k_1 through k_2;

k_1 specifies the lowest frequency index included in the line probing signal; and

k_2 specifies the highest frequency index included in the line probing signal.

61. The modem of claim 60 wherein the phase angles θ_k of the line probing signal are selected to achieve a small peak-to-average ratio for the line probing signal.

62. The modem of claim 60 wherein the phase angles θ_k of the line probing signal are equal to:

$$\theta_k = \pi(k-k_2)^2/(k_2-k_1)$$

63. The modem of claim 46 wherein the receiver further comprises an adaptive filter for providing a desired overall impulse response to a decoder and wherein at least some of the measured characteristics take into account the adaptive filter.

64. The modem of claim 63 wherein the adaptive filter is used in conjuction with trellis precoding.

65. The modem of claim 46 further comprising a transmitter for transmitting information based upon the measured characteristics to the remote device so that the remote device may identify one of said plurality of bit rates based upon said information and then communicate said identified bit rate to the receiver and wherein the selector selects the identified bit rate as said selected bit rate.

66. The modem of claim 46 wherein the line probing processor comprises:

a. a spectrum analyzer for generating discrete spectral representations of the received line probing signal; and

b. a module for computing a nonlinear distortion indicator based upon the discrete spectral representations of the received line probing signal.

67. The modem of claim 46 wherein the receiver comprises a monitor circuit for measuring a power level of the received line probing signal and wherein the measured characteristics includes a quantity derived from the received power level.

68. In a system in which a local modem receives data sent by a remote device over a receiver channel in the form of a first single carrier modulated signal, the modem being capable of receiving the first modulated signal over any one of a first plurality of frequency bands, a method for establishing communication conditions comprising the steps of:

a. sending a first line probing signal from the remote device to the local modem over the receiver channel, said line probing signal simultaneously stimulating more than one of said plurality of frequency bands;

b. receiving the first line probing signal in the local modem;

c. measuring characteristics of the receiver channel based upon the received first line probing signal; and

d. selecting one of the first plurality of frequency bands based upon the measured characteristics of the receiver channel, said selected one of the first plurality of frequency bands to be used for receiving the first modulated signal from the remote device.

69. The method of claim 68 wherein the modem transmits data to the remote device over a transmitter channel by using a second single carrier modulated signal and being capable of sending the second signal over any one of a second plurality of frequency bands, the method further comprising the steps of:

a. sending a second line probing signal from the local modem to the remote device over the transmitter channel;

b. receiving the second line probing signal in the remote device;

c. measuring characteristics of the transmitter channel based upon the received second line probing signal; and

d. selecting one of the second plurality of frequency bands based upon the measured characteristics of the transmission channel, said selected one of the second plurality of frequency bands to be used for sending the second modulated signal to the remote device.

70. In a system in which a local modem receives data sent by a remote device over a receiver channel in the form of a first single carrier modulated signal, the modem being capable of receiving the first modulated signal at any one of a first plurality of bit rates, a method for establishing communication conditions comprising the steps of:

a. sending a first line probing signal from the remote device to the local modem over the receiver channel;

b. receiving the first line probing signal in the local modem;

c. measuring characteristics of the receiver channel based upon the received first line probing signal; and

d. selecting one of the first plurality of bit rates based upon the measured characteristics of the receiver channel, said selected one of the first plurality of bit rates to be used for receiving the first modulated signal from the remote device.

71. The method of claim 70 wherein the modem transmits data to the remote device over a transmitter channel by using a second single carrier modulated signal and is capable of sending the second signal at any one of a second plurality of bit rates, the method further comprising the steps of:

a. sending a second line probing signal from the local modem to the remote device over the transmitter channel;

b. receiving the second line probing signal in the remote device;

c. measuring characteristics of the transmitter channel based upon the received second line probing signal; and

d. selecting one of the second plurality of bit rates based upon the measured characteristics of the transmission channel, said selected one of the second plurality of bit rates to be used for sending the second modulated signal to the remote device.

* * * * *

UNITED STATES PATENT AND TRADEMARK OFFICE

CERTIFICATE OF CORRECTION

PATENT NO. : 5,048,054 Page 1 of 2

DATED : September 10, 1991

INVENTOR(S) : Vedat M. Eyuboglu, et al

It is certified that error appears in the above-identified patent and that said Letters Patent is hereby corrected as shown below:

Col. 4, line 14, after "probing", insert --processors--.

Col. 5, line 60, "f_{66}" should be --f_{Δ}--.

Col. 6, example (9), "f_{66}" should be --f_{Δ}--.

Col. 6, example (10), "f_{66}" should be --f_{Δ}--.

Col. 7, example (11), "f_{66}" should be --f_{Δ}--.

Col. 7, line 9, "Ψ"(kf_{Δ})" should be --$\Phi(kf_{\Delta})$--.

Col. 8, line 57, "h'_{nb}" should be --h'_{n}--.

Col. 9, line 7, delete "L" after "between".

Col. 10, line 33, "describe" should be --described--.

UNITED STATES PATENT AND TRADEMARK OFFICE

CERTIFICATE OF CORRECTION

PATENT NO. : 5,048,054 Page 2 of 2

DATED : September 10, 1991

INVENTOR(S) : Vedat M. Eyuboglu, et al

It is certified that error appears in the above-identified patent and that said Letters Patent is hereby corrected as shown below:

Col. 12, line 29, "Q" should be --Q_1--.

Col. 13, line 41, after "phase", insert --.--.

Col. 14, line 2, after "frequencies", insert --,--.

Signed and Sealed this

Fifth Day of January, 1993

Attest:

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THIS IS TO CERTIFY THAT ANNEXED HERETO IS A TRUE COPY FROM THE RECORDS OF THIS OFFICE OF:

U.S. PATENT: *5,428,641*
ISSUE DATE: *June 27, 1995*

By Authority of the
Under Secretary of Commerce for Intellectual Property and Director of the United States Patent and Trademark Office

M. K. CARTER
Certifying Officer

US005428641A

United States Patent [19]

Long

[11] Patent Number: 5,428,641

[45] Date of Patent: Jun. 27, 1995

[54] **DEVICE AND METHOD FOR UTILIZING ZERO-PADDING CONSTELLATION SWITCHING WITH FRAME MAPPING**

[75] Inventor: Guozhu Long, Canton, Mass.

[73] Assignee: Motorola, Inc., Schaumburg, Ill.

[21] Appl. No.: 97,343

[22] Filed: Jul. 23, 1993

[51] Int. Cl.^{6} H04L 27/04

[52] U.S. Cl. 375/295; 371/43

[58] Field of Search 375/17, 69, 39, 42, 375/94; 371/37.8, 43

[56] **References Cited**

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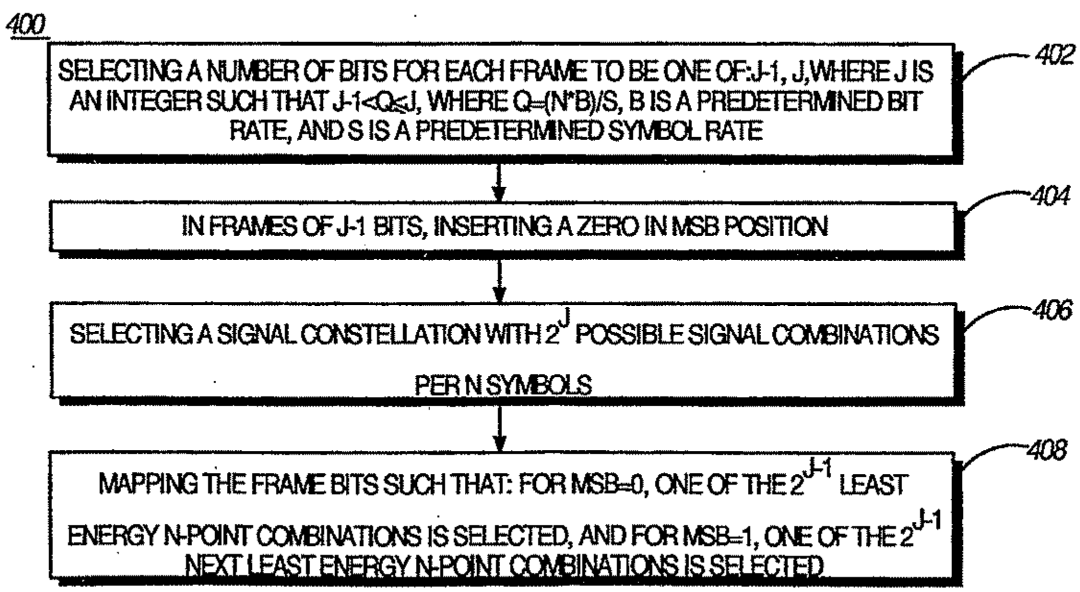
Primary Examiner—Young Tse

Attorney, Agent, or Firm—Darleen J. Stockley

[57] **ABSTRACT**

A device (500) and method (400) for zero-padding constellation switching with frame mapping provides reduced complexity for mapping frames having possibly a fractional number of bits and a predetermined number of symbols while eliminating the usual disadvantages of constellation switching.

8 Claims, 2 Drawing Sheets



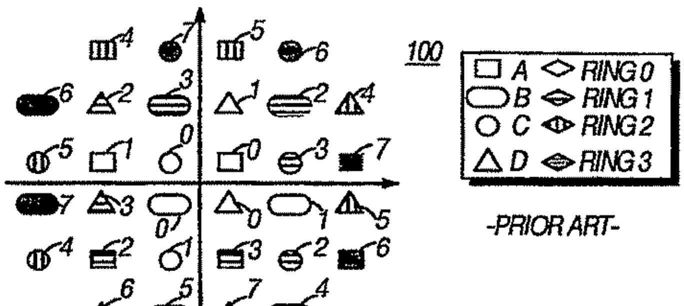


FIG.1

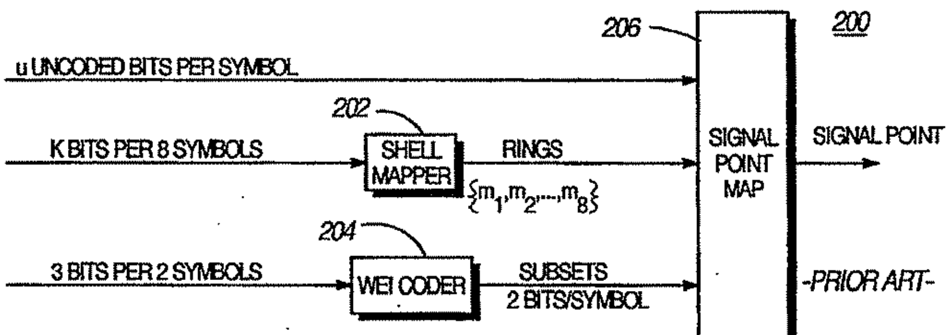


FIG.2

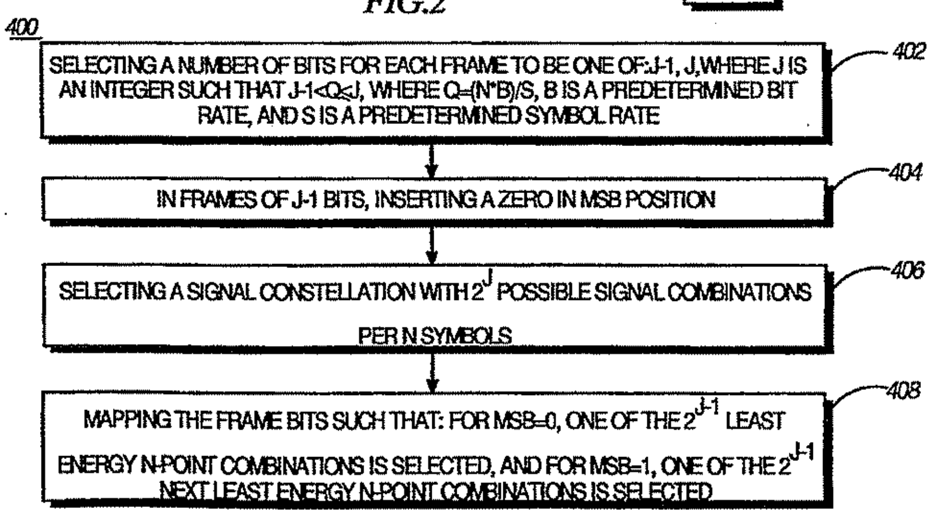
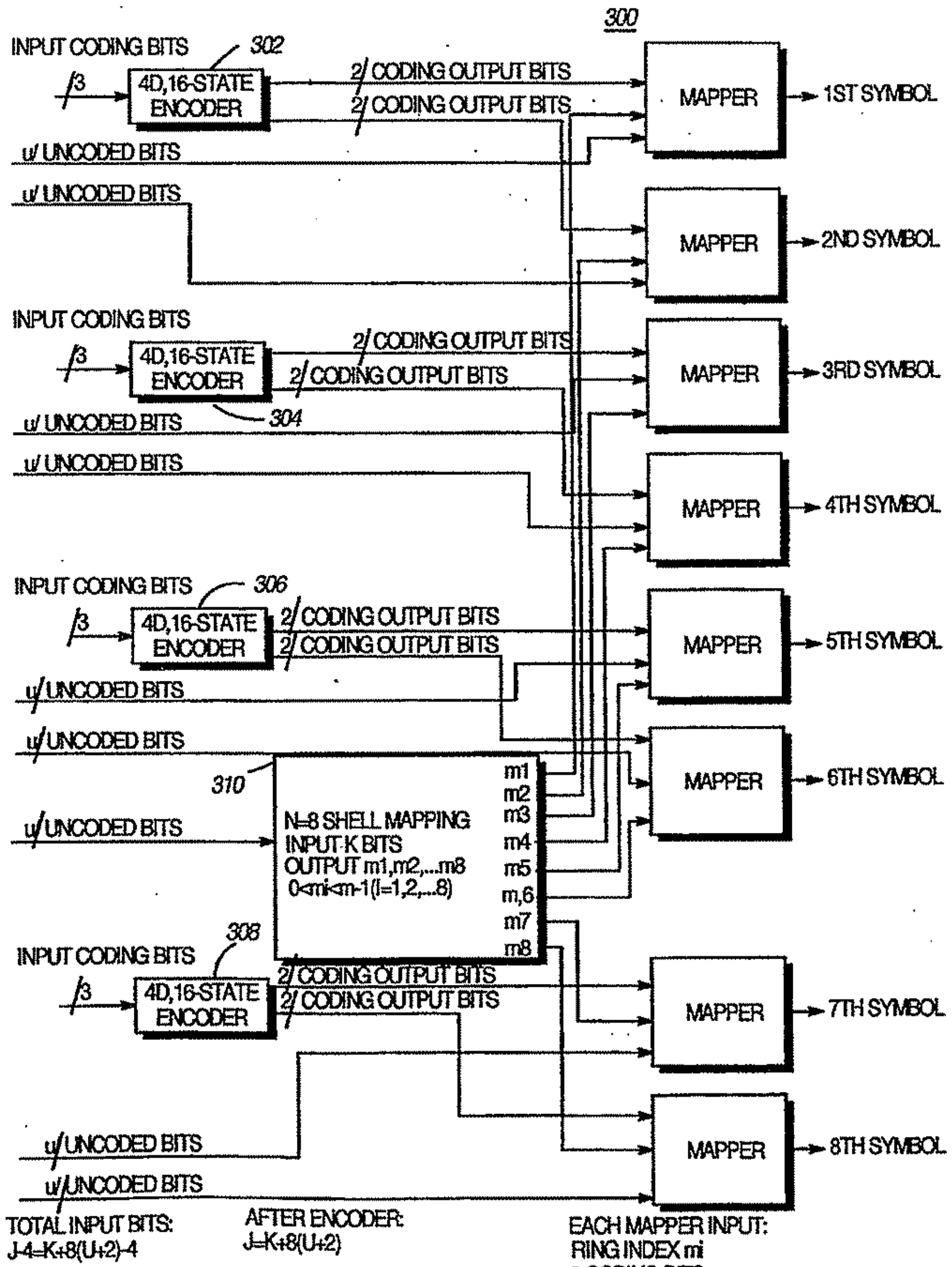


FIG.4

FIG.5



-PRIOR ART-

FIG.3

DEVICE AND METHOD FOR UTILIZING ZERO-PADDING CONSTELLATION SWITCHING WITH FRAME MAPPING

FIELD OF THE INVENTION

The present invention relates generally to digital communication devices and methods, and more particularly to mapping a digital data sequence for transmission in a digital communication system.

BACKGROUND

Modems are typically utilized for transmitting and receiving digital data communication from computer systems. Typically, data is expressed in a binary form. Data is transmitted at a bit rate, e.g., B bits per second, where the bit rate is defined as the number of bits to be transmitted and received, including the actual binary information data rate and a predetermined redundancy needed by coding in a selected system. During transmission, the binary data information is typically transmitted and received in a form of a series of symbols at a symbol rate of S symbols per second. Thus, each symbol contains B/S bits of binary data.

Each symbol can be represented by one of the possible line signal states generated by the modem. Various modulation techniques may be used to convert data into line signal states. For example, in quadrature amplitude modulation (QAM), The line signal states can be represented by a set of complex numbers, namely by a set of points in a two-dimensional signal constellation. For example, for a bit rate B and a symbol rate S, where B/S is an integer D, a signal constellation of size 2^D is needed to represent D bits in each symbol. Thus, if B=12000 bits/second and the symbol rate S is 2400 symbols/second, D=5 bits/symbol, and a 32-point two-dimensional signal constellation is used, providing a scheme for mapping one out of 32 possible complex signal points according to 5 input data bits.

However, for high speed modems, multiple symbol rates and bit rates may be used to facilitate more efficient use of the available channel bandwidth. In such instances, the ratio B/S may not be always an integer.

Where B/S is not an integer, the modem transmits a fractional number of bits per symbol. Various techniques have been used to accomplish transmission of a fractional number of bits per symbol. In the constellation switching technique, for example, where $D-1<B/S<D$, the modem switches between D-1 bits/symbol and D bits/symbol such that, on an average, the modem sends B/S bits per symbol. Namely, the modem switches between a signal constellation with 2^{D-1} points and another signal constellation with 2^D points.

The main disadvantages of constellation switching are the introduction of a variation in constellation size and an increase in peak-to-average power ratio due to the increase in constellation size. This is undesirable in many applications, especially when the communication channel introduces signal-dependent impairments such as harmonic distortion and pulse coded modulation (PCM) noise. In addition, constellation switching often complicates the modem implementation.

In contrast to one symbol by one symbol mapping techniques, frame-mapping techniques may be used.

Since B/S is usually a rational number, there exists a number N such that Q=N*B/S is an integer, i.e., the number of bits in a frame of N symbols is an integer. A frame-mapping algorithm maps the incoming Q bits to N symbols chosen from a signal constellation with a sufficiently large number of points. Compared with the constellation required by symbol-based constelation switching (with 2^D points), frame-mapping techniques usually require a relatively small constellation, and thus a small peak-to-average power ratio. However, such techniques generally introduce complexity where a large N is used. Examples of frame-mapping techniques include modulus conversion and shell mapping.

Since multiple B and S values are often selected, there is often difficulty in selecting a reasonably small N such that Q=N*B/S is always an integer for all possible combinations of B and S. In some instances, a different N may be selected for different combinations of B and S, thereby further complicating the implementation of the mapping algorithm. Another approach is to transmit a fractional number of bits per frame, i.e., using a fixed N in conjunction with a constellation switching among frames. Both approaches provide more implementation complexity.

Hence, there is a need for a frame-mapping device and method which maps data that is transmitted at fractional bits per frame rate such that the implementation difficulties of constellation switching are avoided.

BRIEF DESCRIPTIONS OF THE DRAWINGS

FIG. 1 is an exemplary signal constellation, as is known in the art.

FIG. 2 is a block diagram of a shell mapper as is known in the art.

FIG. 3 is a block diagram of a 4 dimensional/16-state encoder that is configured for shell mapping with N=8, as is known in the art.

FIG. 4 is a flow chart showing the steps of an embodiment of the frame-mapping method of the present invention.

FIG. 5 is a block diagram of a frame-mapping device in accordance with an embodiment of the present invention.

DETAILED DESCRIPTION OF A PREFERRED EMBODIMENT

The present invention provides an advantage for mapping frames of data which would otherwise require constellation switching by utilizing a scheme of selecting first and second frame sizes for data, zero-padding the data for the first frame size such that the first frame is equal to the second frame size, and frame mapping the data.

This scheme is particularly advantageous in that it eliminates the necessity for constellation switching when Q, an average number of bits per frame of N symbols where Q=N,B/S (N is a predetermined integer, B is a number of bits per frame, S is a symbol rate) is expressible in a form of an improper fraction, hereafter referred to as fractional.

Where a frame-mapping technique is used to map a frame of Q=N*B/S bits into N symbols, if Q, the number of bits in a frame of N symbols, is an integer, a frame-mapping algorithm maps the incoming Q bits to N symbols chosen from a signal constellation with a sufficiently large number of points. Examples of frame-mapping techniques include modulus conversion and shell mapping.

In shell mapping, the selected signal constellation is divided into M equal size "rings" each of which contains R signal points. R is typically chosen to be 2^r,

where v is an integer greater than or equal to zero. Among Q bits, K=Q−(N,v) bits, are used for shell mapping, generating N ring indices ranging from 0 to M-1. In each symbol, one of these ring indices is used to choose one of the M rings in the signal constellation, and v bits are used to choose one of the 2^v points within that ring. If a convolutional encoder is employed, among v bits, c bits are coding bits which are the output of the convolutional encoder, and are used to select one of the 2^c subsets, and the remaining u=v−c bits are used to select one of the 2^u points in the chosen subset in the chosen ring. When no convolutional encoder is employed, c=0 and u=v.

For example, for a 4-dimensional 16-state code, c=2, and the constellation is partitioned into 2^c=4 equal-sized subsets A,B,C and D. The constellation is also partitioned into M equal-sized rings, each of which contains 2^{u+c} points, of which 2^u points are from each subset. Ring **0** contains the 2^u lowest energy points from each subset, ring **1** contains the 2^u next lowest energy points from each subset, and so on. An exemplary 32 point signal constellation is shown in FIG. **1**, numeral 100, where L=32, c=2, u=1 and M=4. These 32 points may be divided into 4 rings (shown as rings 0 (unshaded), 1 (shaded with horizontal lines), 2 (shaded with vertical lines), 3 (dotted)), each of which consists of 8 points (indexed 0–1, or 2–3, or 4–5, or 6–7) of which 2 points are from each subset. The energy of the signal point is non-decreasing as the index increases. The collection of all 32 points with index 0 to 7 is approximately within a circle.

The constellation size is $L=M*2^{u+c}$.

Shell mapping uses K=Q−(N*v) bits to generate N ring indices m_i (i=1,2, . . .N) ranging from 0 to M-1. In order to select the N-tuple {m_1, m_2, . . . m_N} using the K bits, first a "cost" is assigned to each ring. Since a main purpose of an efficient mapping scheme is to achieve a low average signal power, the average signal power within a ring is assigned as its "cost". However, for computational convenience, the "cost" is set to the ring index. Thus, the innermost ring has cost **0**, the next ring has cost **1**, and so on. This approximation is quite workable, in particular as the constellation becomes larger. Since costs are additive, the total costs of N symbols is m_1+m_2. . .$+m_N$. Note that there are M^N combinations of N ring indices, but only $2^K \leq M^N$ input K bit combinations. Thus, certain combinations of rings will be excluded. An efficient shell mapping scheme minimizes the cost by selecting the 2^K combinations that have the least total cost. If the input K bits is expressed as a number X, an efficient shell mapping algorithm ensures that as X increases, the corresponding total cost is non-decreasing. As a result, when the most significant bit in the incoming K bits is 0, one of the 2^{K-1} combinations with least total costs will be selected; and when the most significant bit in the incoming K bits is 1, one of the 2^{K-1} combinations with next least total costs will be selected. It is this property that makes this invention possible.

The integer M is chosen such that $2^K \leq M^N$. The smallest integer for M that satisfies this condition is defined as M_{min}.

When $M=M_{min}$, the constellation size is $L=M_{min}*2^{u+c}$, which is in general smaller than the constellation size required by symbol-based constellation switching, thus the peak power is reduced. When M is selected such that $M>M_{min}$, the difference between 2^K and M^N is larger. Interestingly, in this case, the 2^K least energy ring index combinations will have a smaller average total cost. Thus, using the shell mapping algorithm with a larger than the minimum constellation allows reduction of the average cost (shaping gain) at the expense of constellation expansion.

The shell mapping scheme establishes a 1-1 mapping between K binary bits and an N-tuple {m_1, m_2, . . . m_N}. Though this mapping may be done using a table lookup, such a mapping table is often too large to be practical. Instead, some efficient mapping algorithms can be used to achieve the mapping such as the shell mapping technique set forth below.

Since multiple B and S values are often selected, there is often difficulty in selecting a reasonably small N such that Q=N*B/S is always an integer for all possible combinations of B and S. One approach is to fix N for all the cases, but this requires transmitting a fractional number of bits per frame in some of the cases.

Where Q is fractional, an integer value J can always be selected such that $J-1<Q\leq J$. J-1 bits are transmitted in some of the frames and J bits are transmitted in the rest of the frames to achieve an average rate of Q bits per frame. Conventional, this requires a constellation switching among frames. Using the scheme of the present invention, such a constellation switching can be avoided.

The values for J-1 and J represent first and second frame sizes, i.e., number of bits per frame, for incoming data. The key to the invention is that only one signal constellation is selected that is suitable for mapping J bits/frame. That is, a signal constellation has at least 2^J possible signal point combinations per N symbols, i.e., providing transmission of J/N bits per symbol. The frame mapping of the data to the signal points provides that where, in the transmitted data frame, the most significant bit (MSB) of the J bits is equal to zero, one of the 2^{J-1} least energy combinations of N signal points is selected. In frames having only J-1 bits, a zero is inserted at the MSB position such that the J-1 bits frame then contains J bits. The frame mapping scheme is the same for all frames of data, ensuring a less complex implementation than if constellation switching were utilized to achieve mapping for a fractional bit/frame rate. Typical values for N are integers 2 and greater. This scheme is particularly useful for frame sizes of 4 or greater.

Further, where the frame mapping is shell mapping, the present invention is very efficient in providing a minimized constellation expansion and a small average signal power.

To further explain shell mapping, the following exemplary shell mapping technique describes an implementation of shell mapping for a trellis code that is a 4D 16-state Wei code where the frame size is N=8.

An L-point signal constellation B is partitioned into four subsets B_0, B_1, B_2, B_3 according to Ungerboeck set partitioning principles, as is known in the art. B is also partitioned into M equal-sized "rings", where ring 0 contains the 2^u least-energy points from each subset, ring 1 contains the 2^u next lowest energy points, and so forth up to ring M−1. It follows, as above, that $L=M*2^{u+2}$.

In this example, where there are I input bits per frame of 8 symbols, 12 input bits of the I bits are used as the 4D/16-state encoder input, and 8u bits, u an integer, are used as uncoded bits. After encoding, 4 redundant bits are added such that there are a total of J=I+4 bits. As shown in FIG. 2, numeral **200**, a block diagram of a

shell mapper as is known in the art, from the J bits, $K=J-8(u+2)$ bits are used as the input to a shell mapper (202) to select the ring labels $\{m_1, m_2, \ldots, m_8\}$. Two bits per symbol from the Wei coder (204), which includes differential encoding, select a subset B_i, and the remaining u bits per symbol select the signal point in the selected ring of a signal point map (206). To limit complexity, the integer u is selected such that K is as large as possible, but no larger than a predetermined limit K_{max}.

Selection of the ring labels $\{m_1, m_2, \ldots, m_8\}$ is very important in shell mapping. First, a cost is assigned to each ring, typically the cost c(i) of the i'th ring being equal to its label i. The total cost of a combination of rings is the sum of the costs of the individual rings. There are M^8 ring combinations, but only $2K \leq M^8$ input bit combinations. The shell mapper (202) selects the 2K combinations that have the least total cost to minimize the average cost.

For example, where $G_1(z)$ is a generating function of costs of a single ring; i.e.,

$$G_1(z)=\Sigma_i a_1(i)z^i=1+z+z^2+\ldots+z^{M-1}.$$

The coefficient $a_1(i)$ of z^i represents the number of rings that have cost i; namely, $a_1(i)=1$ if $0 \leq i \leq M-1$, and $a_1(i)=0$ otherwise. Similarly, where $G_2j(z)$ be the generating function of costs of 2^j-ring combinations, where $j=1$, 2 or 3; i.e.,

$$G_2j(z)=\Sigma_i a_2j(i)z^i,$$

it is seen that

$$G_8(z)=[G_4(z)]^2=[G_2(z)]^4=[G_1(z)]^8.$$

$z_8(i)$ is defined as the number of eight-ring combinations with cost less than i:

$$z_8(i)=\Sigma 0 \leq j \leq a_8(j).$$

Where $a_2(i)$, $a_4(i)$and $z_8(i)$ are precomputed and stored in memory, the following steps are utilized:

(1) Collect the K input bits and represent them as an integer X, $0 \leq X \leq 2^K-1$.

(2) Find the largest integer C_8, $0 \leq C_8 \leq 8(M-1)$, for which $z_8(C_8) \leq X$. This can be done by a binary search through the table for $z_8(i)$. The value C_8 is the total cost of $(m_1, m_2, \ldots, m_8)$.

(3) Compute the residue $R_8=X-z_8(C_8)$, where $0 \leq R_8 \leq g_8(C_8)-1$. Then use C_8 and R_8 to determine $(m_1, m_2, \ldots, m_8)$ as follows:

First, split C_8 into C_{41} and C_{42}, which are the total costs of (m_1, m_2, m_3, m_4) and (m_5, m_6, m_7, m_8), respectively. Since $C_{41}+C_{42}=C_8$, the following combinations are obtained:

C_{41}	C_{42}	Number of Combinations
0	C8	$a_4(0)a_4(C_8)$
1	C_8-1	$a_4(1)a_4(C_8-1)$
2	C_8-2	$a_4(2)a_4(C_8-2)$
...		
C_8	0	$a_4(C_8)a_4(0)$

It follows that if the convolution terms $a_4(0)a_4(C_8)$, $a_4(1)a_4(C_8-1), \ldots$, are subtracted from R_8 in sequence according to

$$R_8=R_8-a_4(s)a_4(C_8-s),\ s=0, 1, \ldots, C_8,$$

then within at most C_8+1 iterations R_8 will become negative (similar to FIR filtering). When $R_8<0$, the value of s is recorded as C_{41} and C_8-s is denoted as C_{42}. C_{41} and C_{42} represent the costs of the four-tuples (m_1, m_2, m_3, m_4) and (m_5, m_6, m_7, m_8), respectively. Then the last term $a_4(C_{41})a_4(C_{42})$ is added to R_8 to obtain the new residue R_4, $0 \leq R_4 \leq a_4(C_{41})a_4(C_{42})-1$, and R_4 is represented as:

$$R_4=R_{42}a_4(C_{41})+R_{41}$$

where $0 \leq R_{41} \leq a_4(C_{41})-1$ and $0 \leq R_{42} \leq a_4(C_{42})-1$. R_{41} and R_{42} are obtained by dividing R_4 by $a_4(C_{41})$. The quotient is R_{42} and the remainder is $R_{41}=R_4-R_{42}a_4(C_{41})$.

To obtain the four-tuple (m_1, m_2, m_3, m_4) from the residue R_{41} and the cost C_{41}. C_{41} is split into two components, C_{211} and C_{212}, which are the costs of the two-tuples (m_1, m_2) and (m_3, m_4), respectively.

Since $G_4(z)$ is the convolution of $G_2(z)$ with itself, the procedure above can be used; i.e., subtract from R_{41} the convolution terms $a_2(0)a_2(C_{41})$, $a_2(1)a_2(C_{41}-1), \ldots$ in sequence according to

$$R_{41}=R_{41}-a_2(j)a_2(C_{41}-j),\ j=0, 1, \ldots, C_{41},$$

until R_{41} becomes negative (within at most $C_{41}+1$ iterations). Then the values of j and $C_{41}-j$ are recorded as C_{211} and C_{212} the costs of the pairs (m_1, m_2) and (m_3, m_4), respectively. The last term $a_2(C_{211})a_2(C_{212})$ is added back to R_{41} to obtain the new residue R_{21}, $0 \leq R_{21} \leq a_2(C_{211})a_2(C_{212})-1$. R_{21} is then represented in the form

$$R_{21}=R_{212}a_2(C_{211})+R_{211}$$

where $0<R_{211} \leq a_2(C_{211})-1$ and $0 \leq R_{212} \leq a_2(C_{212})-1$. The quantities R_{211} and R_{212} are obtained, as before, by dividing R_{21} by $a_2(C_{211})$.

Determining the rings $(m_1, m_2, m_3, m_4)$is now straightforward; since $a_1(i)=1$, $0 \leq i \leq M-1$, it is clear that:

(a) If $C_{211} \leq M-1$, then $m_1=R_{211}$ and $m_2=C_{211}-R_{211}$.

(b) If $M-1 \leq C_{211} \leq 2(M-1)$, then $m_1=C_{211}-(M-1)+R_{211}$ and $m_2=M-1-R_{211}$.

(c) If $C_{212} \leq M-1$, then $m_3=R_{212}$ and $m_4=C_{212}-R_{212}$.

(d) If $M-1<C_{212} \leq 2(M-1)$, then, $m_3=C_{212}-(M-1)+R_{212}$ and $m_4-M-1-R_{212}$.

Similarly, the four-tuple (m_5, m_6, m_7, m_8) may be obtained from R_{42} and C_{42}.

Thus, the original K bits are uniquely recovered in a receiver from the received eight-tuple $(m_1, m_2, \ldots, m_8)$. The operation of the decoder is easily derived from that of the encoder.

FIG. 3, numeral 300, is a block diagram of a 4-dimensional/16-state encoder that is configured for shell mapping with N=8, as is known in the art. In this embodiment of shell mapping, which is suitable for the shell mapping portion of the present invention, two bits per symbol from the convolutional encoder (302, 304, 306, 308) select the subset A, B C or D. Then the shell mapper (310) utilizes K bits per frame of eight symbols to generate 8 ring indices for selecting rings, and the re-

maining u bits per symbol select the signal point in the appropriate subset within the selected ring.

FIG. 4, numeral **400**, is a flow chart showing the steps of an embodiment of the frame-mapping method of the present invention. The frame-mapping method is utilized for mapping N-symbol frames of data, N a predetermined integer.

The first step (**402**) is selecting a number of bits for each frame to be one of: J-1, J, where J is an integer such that $J-1<Q\leqq J$, where $Q=N*B/S$, B is a predetermined bit rate, and S is a predetermined symbol rate. Then, in frames of J-1 bits, a zero is inserted (**404**) in a most significant bit (MSB) position. A signal constellation is selected (**406**) with at least 2^J possible signal combinations per N symbols. Then, the frame bits are mapped (**408**) such that for MSB-0, one of the 2^{J-1} least energy combinations of N points is selected from the signal constellation such that the averaged energy is minimized. Note that when Q is an integer, Q=J, and all the frames have J bits. Thus, the mapping reduces to conventional frame mapping for integer number of bits per frame.

When using shell mapping, among J bits, K=J-N,v bits are used for shell mapping which generates N ring indices ranging from 0 to M-1. In each symbol, one of these ring indices is used to choose one of the M rings in the signal constellation, and v bits ($v\leqq 0$, an integer) are used to choose one of the 2^v points within that ring. Among v bits, c bits ($c\leqq 0$, an integer) are coding bits which are the output of the convolutional encoder, and are used to select one of the 2^c subsets, and the remaining u=v-c bits($u\leqq 0$, an integer) are used to select one of the 2^u points in the chosen subset in the chosen ring.

Choosing a large K will normally increase the complexity of the shell mapping scheme. Usually, K is limited to less than or equal to K_{max}. The typical values for K_{max} are, for example, 15 or 31. In cases where J is large, K is kept to be less than or equal to K_{max} by selecting a large u. Thus, if B is a two-dimensional signal constellation with $L=M*2^{u+c}$ points, where M is an integer such that $2^{K/N}\leqq M$. M_{min} is defined to be the minimum value for the integer M. M may be selected such that $M>M_{min}$ to obtain more shaping gain and such that $M=M_{min}$ to obtain the smallest constellation size. Selecting a different value for M allows a trade-off between shaping gain and constellation size.

In the embodiment of the method of the present invention, Q is a fractional bits per symbol rate. When the frame-mapping method is shell mapping, the signal constellation is divided into M equal size rings each of which has 2^v ($v\leqq 0$, a predetermined integer) points. In this embodiment, the number of bits in each frame is one of: J-1 and J.

Where J-1 is the number of bits in the frame, K-1 (K=J-N*v, a predetermined integer) bits, together with a zero as the MSB, are utilized for shell mapping, to obtain N ring indices ranging from 0 to M-1 (M an integer) such that an averaged sum of N ring indices obtained in shell mapping is minimized, thereby minimizing average signal power. Where J is the number of bits in the frame, K bits are utilized for shell mapping to obtain N ring indices.

FIG. 5, numeral **500**, is a block diagram of a frame-mapping device in accordance with the embodiment of the present invention. The frame-mapping device maps N-symbol frames of data, N a predetermined integer, such that a fractional number of bits per frame can be transmitted without constellation switching, and includes a frame selector (**502**), a zero insertion unit (**504**), and a signal constellation selector/mapper (**506**). The frame selector (**502**) is operably coupled to receive the data, for selecting a number of bits for each frame to be one of: J-1, J, where J is an integer such that $J-1\leqq Q<J$, where $Q=N*B/S$, B is a predetermined bit rate, and S is a predetermined symbol rate. The zero insertion unit (**504**) is operably coupled to the frame selector (**502**), for, in frames of J-1 bits, inserting a zero in a most significant bit (MSB) position. The signal constellation selector/mapper (**506**) is operably coupled to the zero insertion unit (**504**), for selecting a signal constellation with at least 2^J possible signal combinations per N symbols, and mapping the frame bits such that for MSB=0, one of the 2^{J-1} least energy N-point combinations is chosen from the signal constellation so that the average energy is minimized. The frame-mapping device operates in accordance with the frame-mapping method described above.

Although exemplary embodiments are described above, it will be obvious to those skilled in the art that many alterations and modifications may be made without departing from the invention. Accordingly, it is intended that all such alterations and modifications be included within the spirit and scope of the invention as defined in the appended claims.

I claim:

1. A frame-mapping method for mapping N-symbol frames of data, N a predetermined integer (N>1), such that a fractional number Q of bits per frame can be transmitted without constellation switching, comprising the steps of:

A) selecting a number of bits for each frame to be one of: J-1, J, where J is an integer such that $J-1<Q<J$, where $Q=N*B/S$, B is a predetermined bit rate, and S is a predetermined symbol rate,

B) in frames of J-1 bits, inserting a zero in a most significant bit (MSB) position,

C) selecting a signal constellation with 2^J possible signal combinations per N symbols, and

D) mapping the frame bits such that for MSB=0, one of the 2^{J-1} N-point combinations with least average energy is selected from the signal constellation.

2. The frame-mapping method of claim 1 wherein:

A) the frame-mapping method is shell mapping,

B) the signal constellation is divided into M equal size rings, M an integer, each of which has 2^v ($v\leqq 0$ a predetermined integer) points, and

C) the number of bits in each frame is one of: J-1 and J,

D) where J-1 is the number of bits in the frame, K-1 (K=J-N*v) bits, together with a zero as the MSB, are utilized for shell mapping, to obtain N ring indices ranging from 0 to M-1 (M being an integer) such that an average sum of N ring indices obtained in shell mapping is minimized, thereby minimizing average signal power, and

E) where J is the total number of bits in the frame, K bits are utilized for shell mapping to obtain N ring indices.

3. A frame-mapping device for mapping N-symbol frames of data, N a predetermined integer (N>1), such that a fractional number of bits per frame can be transmitted without constellation switching, comprising:

A) a frame selector, operably coupled to receive the data, for selecting a number of bits for each frame of data to be one of: J-1, J, where J is an integer such that $J-1<Q\leqq J$, where $Q=N*B/S$, B is a

predetermined bit rate, and S is a predetermined symbol rate,

B) a zero insertion unit, operably coupled to the frame selector, for, in frames of J-1 bits, inserting a zero in a most significant bit (MSB) position,

C) a signal constellation selector/mapper, operably coupled to the zero insertion unit, for selecting a signal constellation with at least 2^J possible signal combinations per N symbols, and mapping the frame bits such that for MSB=0, one of 2^{J-1} combinations of N points with least average energy is selected from the signal constellation.

4. The frame-mapping device of claim 3 wherein:

A) the frame-mapping device is a shell mapper,

B) the signal constellation is divided into M equal size rings. M an integer, each of which has 2^v ($v \geqq 0$, a predetermined integer) points, and

C) the number of bits in each frame is one of: J-1 and J,

D) where J-1 is the number of bits in the frame, K-1 (K=J-N*y, an integer) bits, together with a zero as the MSB, are utilized for shell mapping, to obtain N ring indices, the indices ranging from 0 to M-1 (M being an integer) such that an average sum of N ring indices obtained in shell mapping is minimized, thereby minimizing average signal power, and

E) where J is the total number of bits in the frame, K bits are utilized for shell mapping to obtain N ring indices.

5. A frame-mapping method for mapping successive frames of data to groups of N symbols, N a predetermined integer (N>1), such that, on average, a fractional number Q of bits are mappable per frame without constellation switching, comprising the steps of:

A) selecting a number of bits for each frame to be one of: J-1, J, where J is an integer such that J-1<Q<J, according to a predetermined pattern,

B) in frames of J-1 bits, inserting a zero in a most significant bit (MSB) position,

C) selecting a set of 2^J possible combinations of N symbols, where each symbol is chosen from a signal constellation and

D) mapping the frame bits such that for MSB=0, one of the 2^{J-1} possible combinations of N symbols of least average energy is selected from the 2^J possible combinations.

6. The frame-mapping method of claim 5 wherein:

A) the frame-mapping method is shell mapping,

B) the signal constellation is divided into M equal size rings, M an integer, each of which has 2^v ($v>0$ a predetermined integer) points, and

C) where in frames of J-1 bits, K-1 (K=J-N*v) bits, together with a zero as the MSB, are utilized for shell mapping, to obtain N ring indices ranging from 0 to M-1 such that an average sum of N ring indices obtained in shell mapping is kept small, thereby keeping the average signal power small, and

D) where in frames of J bits, K bits are utilized for shell mapping to obtain N ring indices.

7. A frame-mapping device for mapping successive frames of data to groups of N symbols, N a predetermined integer (N>1), such that, on average, a fractional number Q of bits are mappable per frame without constellation switching, comprising:

A) a frame selector, operably coupled to receive the data, for selecting a number of bits for each frame of data to be one of: J-1, J, where J is an integer such that J-1<Q<J, according to a predetermined pattern,

B) a zero insertion unit, operably coupled to the frame selector, for, in frames of J-1 bits, inserting a zero in a most significant bit (MSB) position,

C) a signal constellation selector/mapper, operably coupled to the zero insertion unit, for selecting a set of 2^J possible combinations of N symbols, where each symbol is chosen from a signal constellation, and mapping the frame bits such that for MSB=0, one of 2^{J-1} possible combinations of N symbols of least average energy is selected from the set of 2^J possible combinations.

8. The frame-mapping device of claim 7 wherein:

A) the frame-mapping device is a shell mapper,

B) the number of bits in each frame is one of: J-1 and J,

C) where in frames of J-1 bits, K-1 (K=J*N-v) bits, together with a zero as the MSB, are utilized for shell mapping to obtain N ring indices ranging from 0 to M-1 (M being an integer) such that an average sum of N ring indices obtained in shell mapping is minimized, thereby minimizing an average signal power, and

D) where in frames having a total of J bits, K bits are utilized for shell mapping to obtain N ring indices.

* * * * *

UNITED STATES PATENT AND TRADEMARK OFFICE

CERTIFICATE OF CORRECTION

PATENT NO. : 5,428,641 Page 1 of 2

DATED : Jun. 27, 1995

INVENTOR(S) : Guozhu Long, Canton, Mass.

It is certified that error appears in the above-indentified patent and that said Letters Patent is hereby corrected as shown below:

Column 5:

line16 reads "2K" should read "2^K"

line17 reads "2K" should read "2^K"

line 23 reads "$= \Sigma_i\ a_f(i)z^1$" should read "$= \Sigma_i\ a_f(i)z^i$"

Column 7:

line 23 reads " K= J-N,v " should read " K = J-N*v "

line 27 reads " (v $\leq$ 0, an interger)" should read "(v $\geq$ 0, an interger) "

line 29 reads " (c $\leq$ 0, an interger)" should read "(c $\geq$ 0, an interger) "

line 32 reads " (u $\leq$ 0, an interger)" should read "(u $\geq$ 0, an interger) "

UNITED STATES PATENT AND TRADEMARK OFFICE

CERTIFICATE OF CORRECTION

Page 2 of 2

PATENT NO. : 5,428,641
DATED : Jun. 27, 1995
INVENTOR(S) : Guozhu Long, Canton, Mass.

It is certified that error appears in the above-indentified patent and that said Letters Patent is hereby corrected as shown below:

Column 8:
line 47 reads " (v $\leq$ 0, a predetermined" should read "(v $\geq$ 0, a predetermined "

Column 9
line 21 reads "K=J-N*y" should read "K=J-N*v"

Column 10
line 39 reads "K=J*N-v" should read "K=J-N*v"

Signed and Sealed this
Twenty-eighth Day of January, 1997

Attest:

Bruce Lehman

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Attesting Officer *Commissioner of Patents and Trademarks*

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U.S. PATENT: *5,446,758*
ISSUE DATE: *August 29, 1995*

By Authority of the
Under Secretary of Commerce for Intellectual Property
and Director of the United States Patent and Trademark Office

M. K. CARTER
Certifying Officer

US005446758A

United States Patent [19]

Eyuboglu

[11] **Patent Number: 5,446,758**

[45] **Date of Patent: Aug. 29, 1995**

[54] **DEVICE AND METHOD FOR PRECODING**

[75] Inventor: **M. Vedat Eyuboglu**, Concord, Mass.

[73] Assignee: **Motorola, Inc.**, Schaumburg, Ill.

[21] Appl. No.: **89,319**

[22] Filed: **Jul. 8, 1993**

[51] **Int. Cl.**[6] **H04B 1/10**; H04L 5/12; H04L 27/00; G06F 11/10

[52] **U.S. Cl.** **375/259**; 375/265; 375/340; 371/43

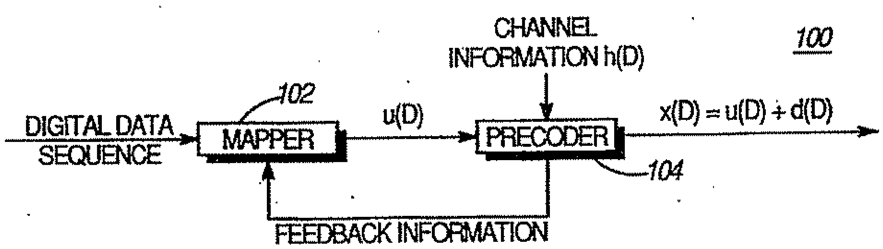
[58] **Field of Search** 375/39, 94, 37; 371/43

Primary Examiner—Edward L. Coles, Sr.
Assistant Examiner—John Ning
Attorney, Agent, or Firm—Darleen J. Stockley

[57] **ABSTRACT**

An improved precoding technique (700) and device (100) allows transmission of a signal point sequence over a channel h(D) to provide efficient data transfer in the presence of intersymbol interference and noise at data rates approaching channel capacity. This improved technique works with trellis-coded modulation and with any signal constellation. Thus, the present invention simplifies shaping and allows signaling at fractional rates without constellation switching. A key advantage of the present invention over prior art is its ability to achieve reduced dither loss by selecting the output components of a mapper based on past components of a channel output sequence.

42 Claims, 3 Drawing Sheets



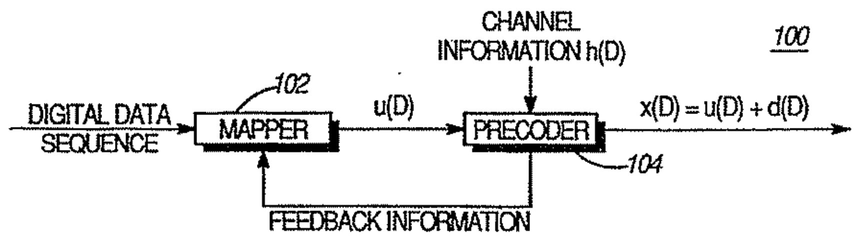


Fig.1

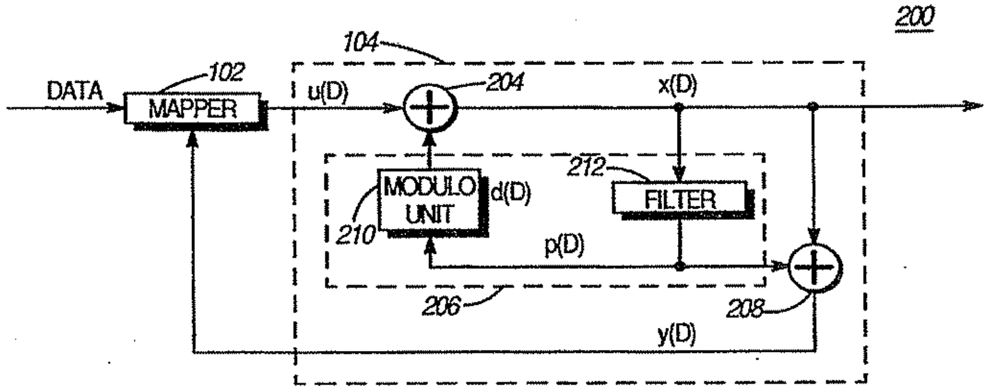


Fig.2

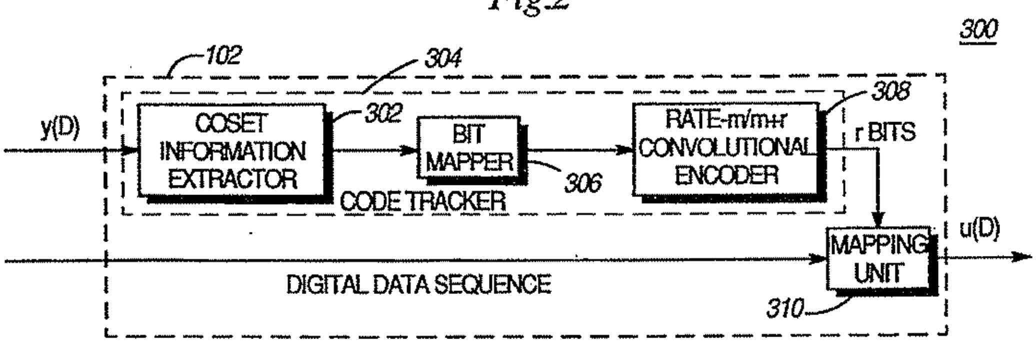


Fig.3

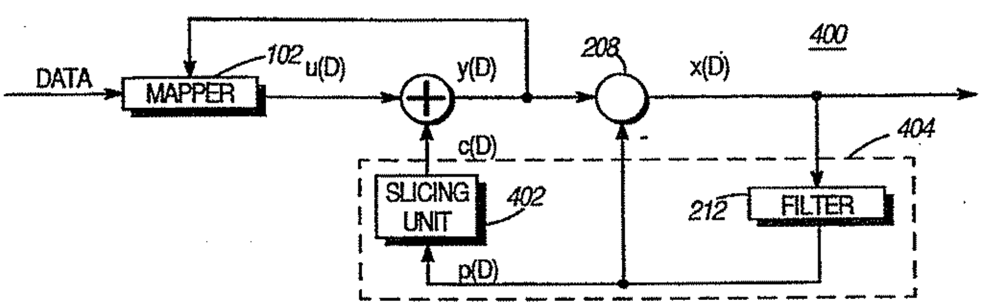


Fig.4

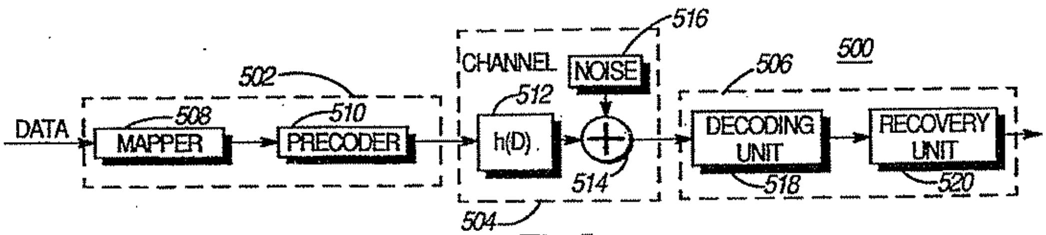


Fig.5

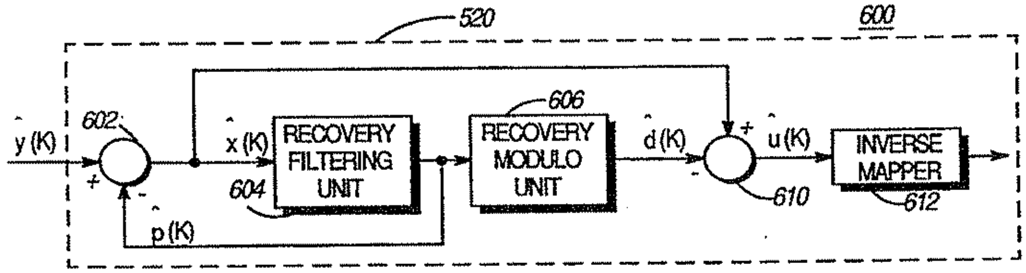


Fig.6

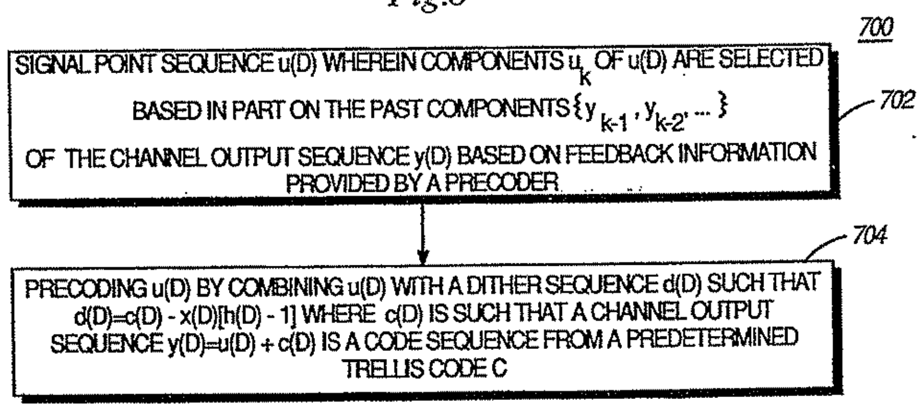


Fig.7

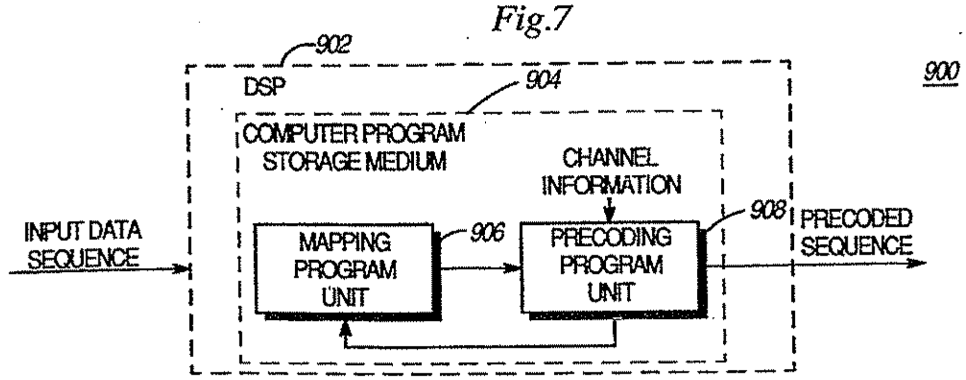
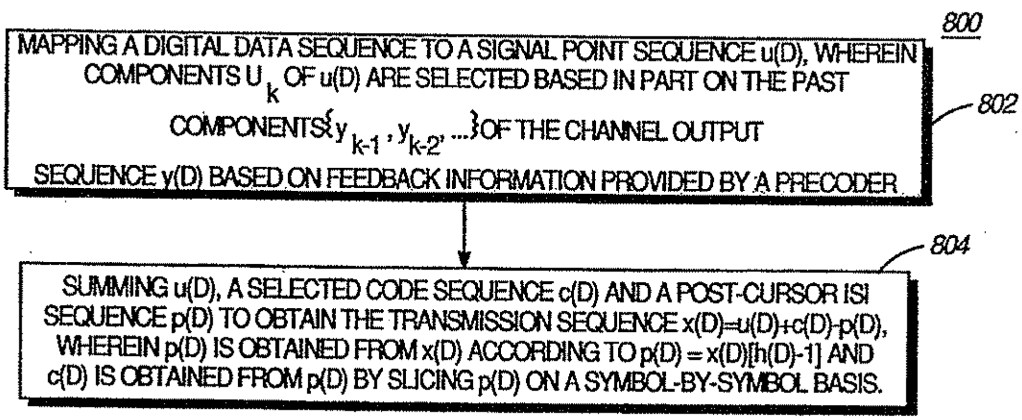


Fig.9


800
MAPPING A DIGITAL DATA SEQUENCE TO A SIGNAL POINT SEQUENCE u(D), WHEREIN COMPONENTS U_k OF u(D) ARE SELECTED BASED IN PART ON THE PAST COMPONENTS $\{y_{k-1}, y_{k-2}, \ldots\}$ OF THE CHANNEL OUTPUT SEQUENCE y(D) BASED ON FEEDBACK INFORMATION PROVIDED BY A PRECODER
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SUMMING u(D), A SELECTED CODE SEQUENCE c(D) AND A POST-CURSOR ISI SEQUENCE p(D) TO OBTAIN THE TRANSMISSION SEQUENCE x(D)=u(D)+c(D)-p(D), WHEREIN p(D) IS OBTAINED FROM x(D) ACCORDING TO p(D) = x(D)[h(D)-1] AND c(D) IS OBTAINED FROM p(D) BY SLICING p(D) ON A SYMBOL-BY-SYMBOL BASIS.

Fig.8

DEVICE AND METHOD FOR PRECODING

FIELD OF THE INVENTION

This invention relates generally to digital communication systems, and more particularly to precoding a digital data sequence for transmission in a digital communication system.

BACKGROUND OF THE INVENTION

It has been shown that on strictly band-limited high-signal-to-noise ratio (SNR) channels with Gaussian noise, digital data may be reliably transmitted at rates approaching channel capacity by using a combination of ideal zero-forcing decision-feedback equalization (DFE) and known coded modulation and constellation shaping techniques designed for ideal channels free of intersymbol interference (ISI). However, ideal DFE is not realizable. Trellis precoding is a realizable combined coding, shaping and equalization technique that achieves the same performance as an ideal DFE along with coding and shaping.

One potential drawback of trellis precoding is that it is effective only for signal constellations whose signal points are uniformly distributed within a space-filling boundary region. Space-filling substantially means that a union of proper non-overlapping translations of the boundary region may cover (tile) the entire space. Stated in another way, the boundary region must be representable as a fundamental region of a lattice, typically referred to as a precoding lattice. To be compatible with known coded modulation techniques, a precoding lattice is typically chosen as a scaled version MZ^2 of a two-dimensional integer lattice Z^2 (where M is a scaling factor) such that the boundary region then has the shape of a square. In certain applications, square signal constellations are not desirable, since they have a higher two-dimensional peak-to-average power ratio (PAR) than constellations with more circular boundaries. More importantly, square constellations are not suitable for representing fractional bits per symbol and require a method known as constellation switching to allow fractional rate transmission, which further increases the two-dimensional PAR. In trellis precoding, it is possible to find precoding lattices whose Voronoi region is more circular than that of a square and which accommodates certain fractional data rates. However, this approach is not very flexible, since it does not uniformly handle all fractional data rates and is more difficult to make invariant to 90° phase rotations, which is an important requirement in certain practical applications. Another drawback of trellis precoding is that to achieve shaping gain, the precoding operation must be combined with shaping operations, which increases the complexity of implementation.

There is a need for a flexible precoding method and device that works with substantially any signal constellation at substantially any data rate and that is implementable independently from constellation shaping while achieving an overall performance that is as close to that of an ideal DFE as possible.

SUMMARY OF THE INVENTION

A device and method are set forth for mapping a digital data sequence into a signal point sequence $x(D)$ for transmission over a channel characterized by a nonideal channel response $h(D)$ using a trellis code C comprising,

- a mapper for mapping the digital data sequence into a signal point sequence $u(D)$ such that a component u_k of $u(D)$ at a given time k is selected based in part on past components $\{y_{k-1}, y_{k-2}, \ldots\}$ of a channel output sequence $y(D)=x(D)h(D)$ based on feedback information provided by a precoder, and
- a precoder for generating said signal point sequence $x(D)$ according to $x(D)=u(D)+d(D)$, wherein $d(D)$ represents a nonzero difference between a selected non-zero sequence $c(D)$ and a postcursor intersymbol interference (ISI) sequence $p(D)$ substantially of a form $p(D)=x(D)[h(D)-1]$, wherein $c(D)$ is selected such that the channel output sequence $y(D)$ is a code sequence in said trellis code C.

BRIEF DESCRIPTIONS OF THE DRAWINGS

FIG. **1** is a block diagram of a device in accordance with the present invention.

FIG. **2** is a more detailed block diagram illustrating a first embodiment of a device in accordance with the present invention.

FIG. **3** is a detailed block diagram of the mapper of FIG. **2**.

FIG. **4** is a block diagram of a second embodiment of a device in accordance with the present invention.

FIG. **5** is a block diagram of a first embodiment of a digital communication system utilizing a device in accordance with the present invention.

FIG. **6** is a block diagram of a recovery unit of the digital communication system of FIG. **5**, showing the recovery unit with more particularity.

FIG. **7** is a flow diagram setting forth steps of one embodiment in accordance with the method of the present invention.

FIG. **8** is a flow diagram setting forth steps of another embodiment in accordance with the method of the present invention.

FIG. **9** is a block diagram of a digital signal processor used for precoding a digital data sequence to obtain a sequence $x(D)$ for transmission over a discrete-time channel with an impulse response $h(D)$ in accordance with the present invention.

DETAILED DESCRIPTION OF A PREFERRED EMBODIMENT

The method and device of the present invention permits precoding a digital data sequence for transmission over a digital communication channel, performing particularly well on channels having severe attenuation distortion. Substantial benefits are obtained by utilizing the present invention: transmission at substantially any desired data rate without constellation switching, transmission with circular signal constellations, simplification of shaping by completely separating shaping from precoding, and reduction of dither loss by selecting the output of the mapper based in part upon the past components of a channel output sequence based on feedback information provided by a precoder.

As illustrated in FIG. **1**, numeral **100**, a device precodes a digital data sequence in accordance with the present invention to provide a complex precoded sequence $x(D)=x_0+x_1D+\ldots$ for transmission over a discrete-time channel unit with a complex impulse response $h(D)=h_0+h_1D+h_2D^2+\ldots$ using a trellis code

C. In this specification, without losing generality, it is assumed that h(D) is monic (i.e., $h_0=1$).

The code C is a 2n-dimensional trellis code, where n is an integer, based on a lattice partition Λ/Λ', where Λ is a preselected lattice and Λ' is a preselected sublattice of Λ, and a rate m/m+r convolutional code. The lattice Λ is the union of 2^r cosets of a so-called time-zero lattice Λ_0 of the trellis code where Λ_0 is a sublattice of Λ and ζ' is a sublattice of Λ_0. If $y(D)=y_0+y_1D+y_2D^2+\ldots$ is a code sequence in the trellis code, and it is represented as a sequence of 2n-dimensional vector components $y_k=(y_{kn}, \ldots, y_{kn+n-1})$, k=0, 1, . . . , then the component y_k at time k will belong to a unique coset of Λ_0 which is determined by the current state S_k at time k.

The device (**100**) includes a mapper (**102**) and a precoder (**104**). A characteristic of the precoding scheme is that the precoded sequence x(D) may be represented by the sum

$$x(D)=u(D)+d(D)$$

where $u(D)=u_0+u_1D+U_2D^2+\ldots$ is a signal point sequence representing the digital data and is provided by the mapper (**102**). The signal points u_i, i=0, 1, 2, . . . are chosen from a two-dimensional signal constellation such that each 2n-dimensional component $U_k=(U_{kn}, \ldots, U_{kn+n-1})$ of U(D) lies on a translate of the lattice Λ. The sequence $d(D)=d_0+d_1D+d_2D^2+\ldots$ is a dither sequence that is generated by the precoder (**104**) according to

$$d(D)=c(D)-p(D)$$

where $p(D)=x(D)[h(D)-1]$ is a post-cursor intersymbol interference (ISI) sequence and $c(D)=c_0+c_1D+c_2D^2+\ldots$ is chosen such that the channel output sequence $y(D)=x(D)h(D)=u(D)+c(D)$ is a code sequence from the trellis code C.

A feature of the present invention is that the components $c_k=(C_{kn}, \ldots, C_{kn+n-1})$ of the sequence c(D) are selected by the precoder (**104**) always from the time-zero lattice Λ_0 of the trellis code C or a sublattice Λ_s thereof, and the components u_k of the sequence u(D) are selected by the mapper (**102**) based in part upon the past values $\{y_{k-1}, y_{k-2}, \ldots\}$ of y(D), such that u_k lies in the coset of Λ_0 in which the component y_k must lie for y(D) to be a valid code sequence in C. The coset is therefore selected based on the state s_k of the code sequence y(D). The digital data determines the signal points from the selected coset as usual. Information about the past values $\{y_{k-1}, y_{k-2}, \ldots\}$ is provided to the mapper (**102**) by the precoder (**104**), which is operably coupled to the mapper (**102**).

The technique of the present invention differs from a technique described in "ISI coder—Combined coding and precoding," AT&T contribution to EIA-TR 30.1, Baltimore, Md., June 1993, in important ways. In the above-referenced technique, the components u_k are always chosen from a translate of the time-zero lattice Λ_0 regardless of the state S_k of y(D), while the components c_k are selected from one of 2^r cosets of Λ_0 depending on the state s_k. The technique of the present invention is logically simpler since the selection of c_k utilizes the same lattice all the time. Furthermore, the method of the present invention may be made transparent to 90 degree rotations (see below), whereas this cannot be achieved completely in the above-referenced technique. Also, in the present invention, the precoder is automatically disabled for ideal channels (i.e., h(D)=1), whereas in the above-referenced technique, a special procedure must be followed on ideal channels.

Typically, the present invention is utilized where the complex impulse response h(D) has no zeroes on the unit circle, or equivalently, when its inverse, 1/h(D), is stable. Therefore, the following embodiments utilize an h(D) that is a canonical response with a stable inverse. Note that h(D) may be an all-zero response such as h(D)=1+0.75 D, an all-pole response such as h(D)=1/(1−0.75 D), or a more general response that includes zeroes and poles. As in earlier precoding techniques, the response h(D) has been determined and is known at the transmitter and the receiver.

FIG. 2, numeral **200**, is a more detailed block diagram of a first embodiment of a device in accordance with the present invention. As in FIG. **1**, the digital data sequence is first mapped in a mapper (**102**) into a signal point sequence u(D) using any combination of known encoding, mapping and shaping techniques. The coset of the time-zero lattice Λ_0 in which the components $u_k=(u_{kn}, u_{kn+n-1}, \ldots, u_{kn+n-1})$ lie is determined based on the past values $\{y_{k-1}, y_{k-2}, \ldots\}$ of y(D) based on feedback information provided by the precoder (**104**). In this embodiment, the precoder comprises a first combining unit (**204**) and a filtering/modulo unit (**206**). The first combining unit (**204**), typically an adder, is operably coupled to receive the input sequence u(D) and a dither sequence d(D) and forms the precoded sequence x(D)=u(D)+d(D). The filtering/modulo unit (**206**) includes a filter (**212**) operably coupled to receive x(D) and is utilized for generating d(D). The filtering/modulo unit (**206**) further includes a modulo unit (**210**) that is operably coupled to receive a post-cursor ISI sequence p(D) from the filtering unit (**212**), where p(D) is obtained by filtering the precoded sequence x(D) that is received from the combining unit (**204**) in accordance with p(D)=x(D)[h(D)−1]. The dither sequence d(D) is given by d(D)=c(D)−p(D), where c(D) is a sequence of 2n-dimensional components c_k chosen from a selected sublattice Λ_s of the time-zero lattice Λ_0. The components c_k are chosen from Λ_s such that the average energy of the dither sequence d(D) is kept small.

A second combining unit (**208**) (typically an adder) combines the precoded sequence x(D) and the post-cursor ISI sequence p(D) to form the channel output sequence y(D) according to y(D)=x(D)+p(D). The sequence y(D) is then fed back to the mapper for coset selection.

Since the components c_k are chosen from a sublattice Λ_s of Λ_0, the channel output components $y_k=u_k+c_k$ will belong to the same coset of Λ_0 as the signal points u_k. Therefore, the channel output sequence y(D) will be a valid code sequence, provided that u_k's are chosen from an allowable coset of the time-zero lattice based on the current state s_k of y(D) in the trellis code.

The modulo unit (**210**) in FIG. 2 is utilized for finding the dither components $d_k=(d_{kn}, d_{kn+1}, \ldots, d_{kn+n-1})$ in a specified fundamental region of the sublattice Λ_s, that are congruent to the negative post-cursor ISI components $-p_k=(-p_{kn}, -p_{kn+1}, \ldots, -p_{kn+n-1})$ modulo the sublattice Λ_s. The sublattice Λ_s is chosen to obtain a good trade-off between complexity and performance. For low complexity, Λ_s may be selected to be an n-fold Cartesian product of a two-dimensional lattice.

The operation of the modulo unit will now be described in more detail with one specific example. Suppose the trellis code is a four-dimensional trellis code that is based on the lattice partition $RZ^4/2D_4$. This code has the time-zero lattice $\Lambda_0=RD_4$, and therefore Λ_s may be chosen as $2Z^4=(2Z^2)^2$ which is a sublattice of RD_4. In this example, the two-dimensional symbols d_{kn+i}, $i=0, 1, \ldots, n-1$, are chosen on a symbol-by-symbol basis from the square region $[-1, 1)x[-1, 1)$ to be congruent to $-p_{kn+i}$ modulo $2Z^2$ (i.e., the real and imaginary parts of d_{kn+i} are congruent to the real and imaginary parts of $-p_{kn+i}$ modulo 2). In some applications, it is important that the precoding scheme be transparent to 90° phase rotations. This is accomplished by using the interval $[-1, 1)$ for positive components of p_k, and $(-1, 1]$ for non-positive components. In this case, the mapper will also include a differential encoder which is well-known in the state-of-the-art (e.g., L. F. Wei, "Trellis-coded modulation using multi-dimensional constellations," IEEE Trans. Inform. Theory, vol. IT-33, pp. 483–501, July 1987).

The energy of the transmitted symbols $S_x=E\{|x_i|^2\}$ will be the sum S_u+S_d of the energies of u(D) and d(D), where E is a statistical expectation. The average energy S_x of the transmitted sequence x(D) will be approximately the same as the energy S_u of the signal sequence u(D) as long as the average dither energy S_d is small. That means, the better the approximation $c(D)\approx p(D)$ is, the smaller will be the increase in average energy due to the dither sequence d(D). This is achieved by choosing the elements of d_k from a fundamental Voronoi region of the sublattice Λ_s. A key advantage of the present invention is that the dither energy is reduced by modifying the sequence u(D) based on the past history of the channel output sequence y(D).

FIG. 3, numeral **300**, shows a more detailed block diagram of the mapper (**102**) which shows a code tracker (**304**) which is utilized to determine the cosets of Λ_0 from which the components u_k are selected. The input to the code tracker (**304**) is the channel output sequence y(D) obtained from the precoder (**104**). The code tracker (**304**) includes a coset information extractor (**302**) for receiving y(D), a bit mapper (**306**) that is operably coupled to the coset information extractor (**302**) and a rate m/m+r convolutional encoder (**308**) that is operably coupled to the mapper. By tracking the cosets of the sublattice Λ' of the trellis code in which successive components y_k lie, it is possible to keep track of the state of the code sequence y(D). This information is used to determine from which of the 2^r cosets of the time-zero lattice Λ_0 the components u_k should be selected. As shown in FIG. 3, it is possible to implement this step by utilizing the coset information extractor (CIE) (**302**) for first extracting the cosets mentioned above using slicing operations, utilizing the bit mapper (**306**) for converting an output of the CIE to m bits (**304**) and then passing those bits through the rate m/m+r convolutional encoder (**308**) to provide r bits which are used together with the digital data symbols by a mapping unit (**310**) to determine the components u_k. These r bits from the convolutional encoder are used to select one of 2^r cosets of Λ_0 in Λ. For example, in the case of a 4D trellis code based on the partition $RZ^4/2D_4$, a rate m/m+1 convolutional encoder is used to produce one extra bit once every two symbols, and this bit is used to select one of two cosets of the time-zero lattice RD^4 in RZ^4, while the data bits select a four-dimensional point from that coset.

FIG. 4 shows an alternative embodiment, numeral **400**, of the present invention which is substantially equivalent to the first embodiment shown in FIG. 2. In this embodiment, a filtering/modulo unit (**404**) includes a slicing unit (**402**) for first selecting the component c_k from the sublattice Λ_s and the filtering unit (**212**) for forming the dither d_k according to $d_k=c_k-p_k$. In this embodiment, the channel output components y_k may be obtained by combining the component u_k with c_k according to $y_k=u_k+c_k$. Information about the sequence y(D) provided by the precoder is used by the code tracker (**310**) in the mapper (**102**) to determine the coset of the time-zero lattice Λ_0 from which the mapper output u_k is selected.

In another example, suppose that again a four-dimensional trellis code based on the partition $RZ^4/2D_4$ is used, but this time the sublattice Λ_s is the time-zero lattice RD_4 itself. First it should be noted that the sublattice RD_4 may be represented as a union of the 4D lattice $2Z^4$ with its coset $2Z^4+(0, 0, 1,1)$. Moreover, the 4D lattice $2Z^4$ can be obtained by taking a Cartesian product of the two-dimensional (2D) lattice $2Z^2$ which consists of all pairs of even integers. Therefore RD_4 is represented as

$$RD_4=(2Z^2\times 2Z^2)U[2Z^2+(1,1)]X[2Z^2+(1,1)],$$

where U represents the union and X represents the Cartesian product. The union of the 2D lattice $2Z^2$ with its coset $2Z^2+(1,1)$ forms the 2D lattice RZ^2.

Therefore, the slicing unit (**410**) may select the 4D symbols $c_k=(c_{2k}, c_{2k+1})$ by selecting, in the even symbol interval 2k, its symbol c_{2k} from RZ^2. If c_{2k} belongs to $2Z^2$ then in the following odd symbol interval, the second symbol c_{2k+1} is selected from the even integer lattice $2Z^2$. If c_{2k} belongs to the coset $2Z^2+(1,1)$, however, then in the next odd symbol interval, the second symbol c_{2k+1} is selected from the coset $2Z^2+(1,1)$. This way it is ensured that the 4D symbol (c_{2k}, c_{2k+1}) will belong to RD_4.

It should be noted that the invention is not limited to criteria that minimize the average dither energy, and any criterion may be used to select the code sequence c(D) as long as the selection of each c_i is based only upon past values x_j, $j<i$, of x(D), and the components c_k belong to the time-zero lattice Λ_0. For example, in certain applications it may be desirable to limit the range of the channel output symbols $y_i=u_i+c_i$. This may be achieved, at the expense of a higher dither energy S_d, by restricting the values of c_i to a certain range.

The above description utilizes an assumption that the channel is characterized by a discrete-time complex impulse response h(D). It is well-known in the state-of-the-art that any discrete time or continuous-time linear channel with additive noise may be represented by a canonical discrete-time equivalent channel with a causal ($h_k=0$, $k<0$), minimum-phase (all zeros outside or on the unit circle), monic ($h_0=1$) impulse response h(D) and additive white noise w(D). A canonical receiver front-end that includes a whitened matched filter and a sampler (in the case of continuous-time channels) operating at a selected symbol rate may be utilized to provide such an equivalent channel. It should be mentioned that in practice, typically, h(D) represents the combined effect of the filters in the transmitter, channel, the receiver, and a sampler. Similarly, w(D) represents the noise after it passes through the receive filters and the sampler. The whitened-matched filter reduces the

strength of the distortion through proper filtering and therein lies the performance advantage of the present invention over conventional linear equalizations.

In practice, when h(D) is an all-zero response, a whitened matched filter may be determined adaptively using standard adaption techniques for decision-feedback equalizers. When it is desired that h(D) be an all-pole filter, then one first determines adaptively an all-zero response h'(D) using the standard methods and then finds $h(D)=1/g'(D)$ using well-known polynomial division techniques, where g'(D) is a finite polynomial approximately equal to $g'(D) \approx 1/h'(D)$.

A first embodiment of a device of the present invention incorporated into a digital communication system is illustrated in the block diagrams of FIG. 5, numeral **500**, wherein at least one of a transmission unit and a receiving unit utilizes the present invention. The said system typically includes at least one of a transmission unit (**502**) and a receiving unit (**506**) wherein the transmission unit has a mapper (**508**) and a precoder (**510**) for transmitting a digital data sequence and a channel (**504**) obtained as described in the above paragraph, operably coupled to the precoder (**510**), for facilitating transmission of the precoded sequence x(D), and the receiving unit (**506**) has a decoding unit (**518**), operably coupled to the channel unit (**504**), for receiving and decoding a received sequence r(D) to provide an estimated output sequence $\hat{y}(D)$, and a recovery unit (**520**), operably coupled to the decoding unit (**518**), for substantially recovering an estimate $\hat{u}(D)$ of the signal point sequence u(D). An estimate of the transmitted digital data sequence is then found from $\hat{u}(D)$ using an inverse map and shaping recovery (if constellation shaping is employed).

The equivalent channel (**504**), represented as set forth above, is substantially represented by a filter (**512**) having a response h(D), for receiving x(D) and producing an output sequence $y(D)=x(D)h(D)$, defined earlier, an additive noise unit (**516**) for providing additive noise, and a combining unit (**514**), typically a summer, operably coupled to the channel filter h(D) (**512**) and to the additive noise unit (**516**).

The decoding unit (**518**) is typically a decoder for the trellis code C, as is known in the art. The decoding unit (**518**), typically receives and decodes a noisy received sequence r(D) which is of a form:

$$\begin{aligned} r(D) &= x(D)h(D) + w(D) \\ &= y(D) + w(D) \\ &= [u(D) + c(D)] + w(D), \end{aligned}$$

to provide an estimate $\hat{y}(D)$ of the channel output sequence $y(D)=x(D)h(D)$, and a recovery unit (**520**), operably coupled to the decoding unit (**518**), substantially recovers an estimate $\hat{u}(D)$ of the input sequence u(D), described more fully below.

As described earlier, the sequence y(D) must be a sequence in the trellis code C. That means that the sequence y(D) may be estimated by a conventional decoder for C, as is known in the art, to provide an estimated output sequence $\hat{y}(D)$.

The recovery unit (**520**), illustrated with more particularity in the block diagram of FIG. 6, numeral **600**, typically includes at least a recovery filtering unit (**604**) substantially the same as the filter (**212**) in the precoder (**104**), operably coupled to receive an estimated sequence $\hat{y}(D)$, for filtering $\hat{y}(D)$ to obtain an estimate $\hat{p}(D)$ of the post-cursor ISI sequence p(D), substantially of a form $\hat{p}(D)=\hat{y}(D)/\{1-1/h(D)\}$ and for providing $\hat{p}(D)$ as a feedback signal to obtain $\hat{x}(D)$ from $\hat{y}(D)$ according to $\hat{x}(D)=\hat{y}(D)-\hat{p}(D)$, a recovery modulo unit (**606**), operably coupled to the recovery filtering unit (**604**), for finding the estimated dither sequence $\hat{d}(D)$, as the sequence whose components $\hat{d}_k=(\hat{d}_{kn}, \hat{d}_{kn+1}, \ldots, \hat{d}_{kn+n-1})$ lie in a specified fundamental region of the sublattice Λ_s, and are congruent to the negative post-cursor ISI components $-\hat{p}_k=(-\hat{p}_{kn}, -\hat{p}_{kn+1}, \ldots, -\hat{p}_{kn+n-1})$ modulo the sublattice Λ_s in a manner that is substantially the same as that used in the precoding unit at the transmitter, and a recovery combining unit (**606**), operably coupled to the recovery modulo unit (**604**) and to the estimator combiner (**602**) for receiving the estimated sequence $\hat{x}(D)$ of the precoded sequence x(D), for substantially determining a difference between the sequence $\hat{x}(D)$ and the sequence $\hat{d}(D)$ to obtain the estimate $\hat{u}(D)$ of the original input sequence u(D). As long as there are no decision errors ($\hat{y}(D)=y(D)$),and the operations in the transmitter and receiver are substantially symmetrical, the original sequence u(k) will be correctly recovered. Other equivalent implementations of the recovery circuit are also possible.

To summarize, the recovery filtering unit (**604**) is utilized to reconstruct an estimate $\hat{p}(D)$ of a post-cursor ISI variable p(D), then the recovery modulo unit (**604**) is utilized to determine a dither sequence $\hat{d}(D)$ that substantially correlates with d(D) in the corresponding device (**100**) at the transmitter, and then utilizes the recovery combining unit (**608**) to provide $\hat{u}(D)=\hat{x}(D)-\hat{d}(D)$.

Of course, there will be occasional errors in $\hat{y}(D)$ due to channel noise, and these may lead to error propagation. However, since 1/h(D) is stable, the error propagation in the filter $1-1/h(D)$ will never be catastrophic. Moreover, if h(D) is an all-pole response of order p (where m is a selected integer), then error propagation will be strictly limited to at most p symbols.

Where an estimate $\hat{u}_i$, an i'th variable of the recovered sequence u(D), falls outside the allowed range of a i'th variable u_i of the input sequence u(D), such a range violation indicates that a decision error has occurred either in the current symbol y_i, or c_i is in error because of an error in some recent symbol y_{j-i}, $i>0$. When such range violations are detected, one may try to correct the violations by adjusting the estimates y_i or y_{j-i}. Thus, by monitoring the range violations, some degree of error correction is achieved. Such an error detection capability may also be useful for monitoring the performance of the transmission system.

Thus, a digital communications receiver may be utilized in accordance with the present invention for receiving a digital data sequence that was mapped into a precoded sequence x(D) and transmitted over a channel characterized by a nonideal response h(D) using a trellis code C, providing a received sequence r(D), where a receiver includes at least a decoding unit, operably coupled to receive r(D), for decoding the received transmission sequence r(D) to provide an estimated output sequence $\hat{y}(D)$, and a recovery unit, operably coupled to the decoding unit, for substantially recovering an estimated sequence $\hat{u}(D)$ for a sequence u(D) for a transmitted signal point sequence x(D) which is generated according to $x(D)=u(D)+d(D)$, wherein u(D) is a signal point sequence which uniquely represents said digital data sequence and wherein the coset of the time-

zero lattice Λ_0 in which 2n-dimensional components u_k lie depends on the state of the channel output sequence y(D), and d(D) represents a nonzero difference between a selected sequence c(D) whose components c_k are selected from a sublattice Λ_s of the time-zero lattice Λ_0 of the trellis code and a post-cursor intersymbol interference (ISI) sequence p(D) substantially of a form p(D)=x(D)[h(D)−1], such that c(D) is selected based only upon p(D).

As illustrated in FIG. 6, one embodiment of the recovery means utilizes an estimator combining unit (602) that is operably coupled to receive the estimated output sequence $\hat{y}(D)$ and $\hat{p}(D)$, a recovery filtering unit (604), operably coupled to receive the estimated output sequence $\hat{x}(D)$, for providing an estimated post-cursor intersymbol interference (ISI) sequence $\hat{p}(D)$, a recovery modulo unit (606), operably coupled to the recovery filtering means, for providing an estimated dither sequence $\hat{d}(D)$ that substantially correlates with d(D) utilized for providing the transmission sequence x(D), a third combining unit (610) (typically a summer), operably coupled to receive the estimated precoded sequence $\hat{x}(D)$ and to the output $\hat{d}(D)$ of the recovery modulo means, for determining the estimated sequence $\hat{u}(D)$, substantially of a form $\hat{u}(D)=\hat{x}(D)-\hat{d}(D)$, and an inverse mapper (612), operably coupled to the third combining means, for inverse mapping the estimated sequence $\hat{u}(D)$ to provide a recovered digital data sequence substantially equal to the transmitted digital data sequence.

In addition the digital communications receiver may be selected such that the decoding unit (518) further includes a reduced complexity sequence estimator unit that utilizes a correlation between successive symbols y_i. In one implementation, the reduced complexity sequence estimator unit utilizes a sequence estimator having a reduced number of states that are determined utilizing state merging techniques for reduced-state sequence estimation (RSSE).

Where desired, the recovery unit may be selected to include a range violation determiner unit. When an i'th variable $\hat{u}_i$ of the recovered sequence $\hat{u}(D)$ is outside a certain range (a range violation), this unit adjusts at least one of an estimate $\hat{y}_i$ and a past estimate $\hat{y}_{k-j}$ (where j is a positive integer) to substantially correct the range violation.

FIG. 7, numeral 700, sets forth a flow chart illustrating steps in accordance with the method of the present invention for precoding a stream of signal points for transmission in a digital communication system. The method provides for precoding a digital data sequence to generate a sequence x(D) for transmission over a discrete-time channel with an impulse response h(D) using a trellis code C. A stream of signal points u(D) is transmitted as x(D)=u(D)+d(D), where d(D) is a dither sequence of a form d(D)=c(D)−p(D), where p(D)=x(D)[h(D)−1]represents a post-cursor intersymbol interference (ISI), and c(D) is a sequence whose components c_k belong to a sublattice Λ_s of the time-zero lattice Λ_0 of a trellis code C, and c(D) is obtained based only upon p(D). The sequence u(D) uniquely represents the digital data sequence, and its components u_k are selected based in part on past components $\{y_{k-1}, y_{k-2}, \ldots\}$ of the channel output sequence y(D)=x(D)h(D), based on feedback information provided by the precoder, such that y(D) is a code sequence in the trellis code.

In one embodiment, illustrated in FIG. 7, numeral 700, the method for mapping a digital data sequence into a signal point sequence x(D) for transmission over a channel characterized by a nonideal channel response h(D) using a trellis code C includes the steps of: (1) mapping the digital data sequence to a signal point sequence u(D) (702), and (2) precoding the signal point sequence u(D) by combining u(D) with a dither sequence d(D) (704) such that d(D) is of a form: d(D)=c(D)−x(D)[h(D)−1], where c(D) is selected such that a channel output sequence y(D)=u(D)+c(D) is a code sequence from the trellis code C. The components u_k of u(D) are selected based in part on past components $\{y_{k-1}, y_{k-2}, \ldots\}$ of the channel output sequence y(D)=x(D)h(D). The components c_k of c(D) are selected from a sublattice Λ_s of the time-zero lattice Λ_0 of the trellis code. The sequences are as described above.

In another embodiment, illustrated in FIG. 8, numeral 800, the method comprises the steps of mapping the digital data sequence to a signal point sequence u(D) (802), summing u(D), a selected code sequence c(D) and a post-cursor ISI sequence p(D) to obtain the transmission sequence x(D)=u(D)+c(D)−p(D) (804), filtering x(D) to obtain (806) p(D) substantially of a form:

$$p(D)=x(D)[h(D)-1],$$

and slicing p(D) on a symbol-by-symbol basis to obtain the sequence c(D). Further modifications of the method may be utilized in accordance with the modifications described more fully above for the device of the present invention. Again, the components u_k of u(D) are selected based in part on past components $\{y_{k-1}, y_{k-2}, \ldots\}$ of the channel output sequence y(D)=x(D)h(D). The components c_k of c(D) are selected from a sublattice Λ_s of the time-zero lattice Λ_0 of the trellis code.

The present invention may be implemented in a digital communication system, illustrated in FIG. 9, numeral 900, where a digital signal processor (902) is utilized to precode a digital data sequence to obtain a sequence x(D) for transmission over a discrete-time channel with an impulse response h(D). The processor typically includes a program storage medium (904) having a computer program to be executed by the digital signal processor, the program including a mapping program (906) for mapping the digital data sequence into a signal point sequence u(D) based in part upon the past values of a channel output sequence y(D) based on information provided by a precoding program and a precoding program (908) for utilizing the u(D) to generate a sequence x(D) wherein x(D) may be represented as the sum u(D)+d(D) of the stream of signal points u(D) which uniquely represents the digital data sequence and is also chosen based on the state of channel output sequence y(D) and a dither sequence d(D)=c(D)−p(D), where c(D) is a sequence from a sublattice Λ_s of the time-zero lattice Λ_0 of the trellis code C and where p(D) represents a post-cursor intersymbol interference (ISI) sequence of a form p(D)=x(D){h(D)−1}. The code sequence c(D) is determined based upon only the post-cursor ISI sequence p(D). The components u_k of u(D) are selected based in part on past components $\{y_{k-1}, y_{k-2}, \ldots\}$ of the channel output sequence y(D)=x(D)h(D). The components c_k of c(D) are selected from a sublattice Λ_s of the time-zero lattice Λ_0 of the trellis code. Further description of the operation of the processor follows that described above.

The processor typically includes a computer program storage medium having a computer program to be executed by the digital signal processor where the computer program includes a mapping program for mapping the digital data sequence into a signal point sequence u(D) based in part upon the past values of a channel output sequence y(D) based on feedback information provided by a precoding program, and a precoding program for selecting said signal point sequence x(D) from a subset of all possible signal point sequences that are of a form x(D)=u(D)+d(D), wherein d(D) represents a nonzero difference between a selected nonzero sequence c(D) and a postcursor intersymbol interference (ISI) sequence p(D) substantially of a form p(D)=x(D)[h(D)−1], where c(D) is selected based upon p(D) such that the channel output sequence y(D)=u(D)+c(D) is a code sequence from a translate of said trellis code C. In one embodiment, the precoding program includes first combining instruction(s) for combining the input sequence u(D) and a dither sequence d(D) to generate the precoded sequence x(D)=u(D)+d(D), d(D) generating/p(D) generating instructions for generating d(D) and for providing a post-cursor ISI sequence p(D), second combining instructions for generating a sequence y(D)=x(D)+p(D).

The present invention relies on past channel output signals to remove a dither sequence d(D) that is added to an input sequence u(D) at the transmitter to form a transmitted sequence, x(D)=u(D)+d(D), the dither sequence being substantially a difference between a post-cursor intersymbol interference p(D) and an appropriate sequence, c(D), from a sublattice Λ_s of the time-zero lattice Λ_0 of a trellis code. The sequence u(D) is chosen based in part on past components $\{y_{k-1}, y_{k-2}, \ldots\}$ of the channel output sequence y(D)=x(D)h(D) based on feedback information provided by the precoder. The coset of the time-zero lattice Λ_0 in which successive elements of u(D) should lie are determined based on the past values $\{y_{k-1}, y_{k-2}, \ldots\}$.

The present invention may be utilized with virtually any signaling method and at any data rate. Further, the present invention may be utilized independently of constellation shaping techniques (e.g., shell mapping, "Signal mapping and shaping for V.fast," Motorola contribution D196, CCITT Study Group XVIII, Geneva, June 1992); that means u(D) may represent an already shaped sequence whose signal points have a nonuniform Gaussian-like probability distribution.

In the present invention, the dither sequence may increase the average transmit energy. Since in practice, the average transmit energy must be kept constant, the signal x(D) must be scaled down to maintain the same average energy. The increase in the average transmit energy is referred to herein as a dithering loss.

Although several exemplary embodiments are described above, it will be obvious to those skilled in the art that many alterations and modifications may be made without departing from the invention. For example, even though primarily trellis codes are described above, the method may be used with block or lattice codes as well. It may also be used with selected multilevel trellis codes. Also, although two-dimensional (passband, quadrature) transmission systems are emphasized, the methods may also be applied to one-dimensional (baseband) or higher-dimensional (parallel channels) transmission systems. Further, the invention may be utilized with trellis codes whose dimensionality is odd. Although the description above emphasizes channel responses h(D) that are monic, the invention may also be applied to more general channel responses with $h_0 \neq 1$, by either scaling the channel response to make it monic, or by appropriately scaling the variables of the precoding system. All such implementations are considered substantially equivalent to the present invention.

Accordingly, it is intended that all such alterations and modifications be included within the spirit and scope of the invention as defined in the appended claims.

I claim:

1. A device for mapping an input digital data sequence into an output signal point sequence x(D) for transmission over a channel characterized by a nonideal channel response h(D) using a trellis code C comprising:

a mapper for mapping the digital data sequence into a signal point sequence u(D) such that the components u_k of u(D), where k is a time index, are selected based in part on past components $\{y_{k-1}, y_{k-2}, \ldots\}$ of a channel output sequence y(D)=x(D)h(D) which are obtained based on feedback information provided by a precoder,

a precoder for generating said output signal point sequence x(D) according to x(D)=u(D)+d(D), wherein d(D) represents a nonzero difference between a selected sequence c(D) and a postcursor intersymbol interference (ISI) sequence p(D) substantially of a form p(D)=x(D)[h(D)−1], wherein c(D)is selected such that the channel output sequence y(D) is a code sequence in said trellis code C.

2. The device of claim 1 wherein the components c_k of c(D) are selected from a time-zero lattice Λ_0 of said trellis code or a sublattice $\Lambda_s$thereof.

3. The device of claim 1 wherein said components u_k of u(D) are selected based on the state S_k of the channel output sequence y(D) in said trellis code.

4. The device of claim 3 wherein said components u_k of u(D) at time k are selected from one of the cosets of the time-zero lattice Λ_0 based on the state s_k of the channel output sequence y(D) in said trellis code.

5. The device of claim 4 wherein said cosets of the time-zero lattice are determined by utilizing a convolutional encoder for the said trellis encoder.

6. The device of claim 1 wherein said trellis code is a four-dimensional trellis code.

7. The device of claim 6 wherein the trellis code is based on the lattice partition $RZ^4/2D_4$ and its time-zero lattice is RD_4.

8. The device of claim 7, wherein said mapper further includes a differential encoder.

9. The device of claim 8, wherein said mapper further includes constellation shaping.

10. The device of claim 9, wherein said constellation shaping is achieved using shell mapping.

11. The device of claim 6 wherein a slicing means provides the selected sequence c(D) by selecting the symbols c_{2k} and c_{2k+1} from the integer lattice $2Z^2$.

12. The device of claim 6 wherein a slicing means provides the selected sequence c(D) by selecting a symbol C_{2k} from RZ^2 in a first symbol interval, a symbol c_{2k+1} from either $2Z^2$ or its coset $2Z^2+(1,0)$ in a second symbol interval, based on c_{2k}, where k is a time index.

13. The device of claim 6 wherein, the selection of c_k further includes constraints to limit a range of the symbols y_i of the channel output sequence y(D).

14. The device of claim 1 wherein the precoder comprises at least:

first combining means, operably coupled to the mapper, and to a modulo means, for combining u(D) and at least the dither sequence d(D) to provide a precoded sequence x(D),

wherein the modulo means is operably coupled to a filter, and is utilized for finding the dither sequence d(D) based on the post-cursor ISI sequence p(D),

wherein the filter is operably coupled to the first combining means, and is utilized for extracting the post-cursor ISI sequence p(D) from the transmission sequence x(D).

15. The device of claim 14 wherein the first combining means is a summer.

16. The device of claim 14 wherein the precoder further includes second combining means, operably coupled to the first combining means and to the filter, for obtaining the channel output sequence y(D) substantially of a form y(D)=x(D)+p(D).

17. The device of claim 1 wherein the precoding unit comprises at least:

first combining means, operably coupled to the mapping means and to a slicing means, for combining u(D) and at least the sequence c(D) to provide the channel output sequence y(D),

wherein the slicing means is operably coupled to a filter, and is utilized for slicing the post-cursor sequence p(D) to a sequence of signal points c(D) whose components C_k are selected from a time-zero lattice Λ_0 of the trellis code C or a sublattice Λ_s thereof,

a second combining means, operably coupled to the first combining means and the filter, for combining the channel output sequence y(D) and the post-cursor ISI sequence p(D) to form the precoded sequence x(D), and

wherein the filter is operably coupled to the second combining means, and is utilized for extracting the post-cursor ISI sequence p(D) from the transmission sequence x(D).

18. The device of claim 17 wherein the first combining means is a summer.

19. The device of claim 17 wherein the second combining means is a summer.

20. The device of claim 1 wherein the trellis code C is a block code.

21. A digital communications receiver for receiving a digital data sequence that was mapped into a signal point sequence x(D) and transmitted over a channel characterized by a nonideal response h(D) using a trellis code C, providing a received sequence r(D), comprising at least:

decoding means, operably coupled to receive r(D), for decoding the received transmission sequence r(D) to provide an estimated output sequence $\hat{y}$(D), and

recovery means, operably coupled to the decoding means, for substantially recovering an estimated sequence $\hat{u}$(D) for a sequence u(D) representing said digital data sequence by first recovering an estimate of the transmitted signal point sequence x(D) generated by a precoder according to x(D)=u(D)+d(D), wherein d(D) represents a nonzero difference between a selected non-zero sequence c(D) and a postcursor intersymbol interference (ISI) sequence p(D) substantially of a form p(D)=x(D)[h(D)−1], wherein c(D) is selected such that the channel output sequence y(D)=x(D)h(D) is a code sequence in said trellis code C and u(D) is selected such that a component u_k at a given time k is selected based in part on past components $\{y_{k-1}, y_{k-2}, \ldots\}$ of a channel output sequence y(D) based on feedback information provided by the precoder, where $u(D)=u_0+u_1D+u_2D^2+, \ldots$ is a signal point sequence representing the digital data provided by a mapper and transmitted to the digital communications receiver.

22. The digital communications receiver of claim 21 wherein the recovery means includes at least:

recovery filtering means, operably coupled to receive the estimated output sequence $\hat{y}$(D), for providing an estimated post-cursor intersymbol interference (ISI) sequence $\hat{p}$(D),

recovery slicing means, operably coupled to the recovery filtering means, for providing an estimated nonzero dither sequence $\hat{d}$(D) that substantially correlates with d(D) utilized for providing the transmission sequence x(D),

third combining means, operably coupled to receive the estimated output sequence $\hat{y}$(D) and to the recovery slicing means, for determining the estimated sequence $\hat{u}$(D), substantially of a form $\hat{u}(D)=\hat{y}(D)-\hat{p}(D)-\hat{d}(D)$, and

an inverse mapping means, operably coupled to the third combining means, for inverse mapping the estimated sequence $\hat{u}$(D) to provide a recovered digital data sequence.

23. The digital communications receiver of claim 21 wherein the nonideal response h(D) represents the impulse response of a noise prediction filter.

24. The digital communications receiver of claim 21 wherein the decoding means further includes a reduced complexity sequence estimator means that utilizes a correlation between successive symbols y_i.

25. The digital communications receiver of claim 24 wherein the reduced complexity sequence estimator means utilizes a sequence estimator having a reduced number of states that are determined utilizing state merging techniques for reduced-state sequence estimation (RSSE).

26. A method for mapping an input digital data sequence into an output signal point sequence x(D) for transmission over a channel characterized by a nonideal channel response h(D) using a trellis code C comprising the steps of:

mapping the digital data sequence into a signal point sequence u(D) such that a component u_k of u(D) at a given time k is selected based in part on past components $\{y_{k-1}, y_{k-2}, \ldots\}$ of a channel output sequence y(D)=x(D)h(D) based on feedback information provided by a precoder,

generating, by the precoder, said signal point sequence x(D) according to x(D)=u(D)+d(D), wherein d(D) represents a nonzero difference between a selected non-zero sequence c(D) and a postcursor intersymbol interference (ISI) sequence p(D) substantially of a form p(D)=x(D)[h(D)−1], wherein c(D) is selected such that the channel output sequence y(D) is a code sequence in said trellis code C.

27. The method of claim 26 wherein the components C_k of c(D) are selected from a time-zero lattice Λ_0 of said trellis code or a sublattice Λ_s thereof.

28. The method of claim 26 wherein said components u_k of u(D) are selected based on the state s_k of the channel output sequence y(D) in said trellis code.

29. The method of claim 28 wherein said components u_k of u(D) at time k are selected from one of the cosets of the time-zero lattice Λ_0 based on the state S_k of the channel output sequence y(D) in said trellis code.

30. The method of claim 29 wherein said cosets of the time-zero lattice are determined by utilizing a convolutional encoder for the said trellis encoder.

31. The method of claim 26 wherein said trellis code is a four-dimensional trellis code.

32. The method of claim 31 wherein the trellis code is based on the lattice partition $RZ^4/2D_4$ and its time-zero lattice is RD_4.

33. The method of claim 32, wherein said mapper further includes a differential encoder.

34. The method of claim 33, wherein said mapper further includes constellation shaping.

35. The method of claim 34, wherein said constellation shaping is achieved using shell mapping.

36. A digital signal processor having at least a plurality of registers and an arithmetic logic unit, for use in a digital communication system to precode a digital data sequence into a signal point sequence x(D) for transmission over a discrete-time channel with a impulse response h(D) using a trellis code C, the processor having a device comprising:

mapping means for mapping, using the plurality of registers and the arithmetic logic unit, the digital data sequence into a signal point sequence u(D) such that a component u_k of u(D) at a given time k is selected based in part on past components $\{y_{k-1}, y_{k-2}, \ldots\}$ of a channel output sequence y(D)=x(D)h(D) based on feedback information,

precoding means for generating, using the plurality of registers and the arithmetic logic unit, said signal point sequence x(D) according to x(D)=u(D)+d(D), wherein d(D) represents a nonzero difference between a selected non-zero sequence c(D) and a postcursor intersymbol interference (ISI) sequence p(D) substantially of a form p(D)=x(D)[h(D)−1], wherein c(D) is selected such that the channel output sequence y(D) is a code sequence in said trellis code C.

37. The signal processor of claim 36 wherein the components c_k of c(D) are selected from a time-zero lattice Λ_0 of said trellis code or a sublattice Λ_s thereof.

38. The signal processor of claim 36 wherein said components u_k of u(D) are selected based on the state s_k of the channel output sequence y(D) in said trellis code.

39. A digital communication system for precoding a digital data sequence into a signal point sequence x(D) for transmission over a discrete-time channel with a impulse response h(D) using a trellis code C, comprising at least one of:

a transmission unit and

a receiving unit,

wherein the transmission unit having a digital signal processor with at least a plurality of registers and an arithmetic logic unit, comprises:

a mapper for mapping, using the plurality of registers and the arithmetic logic unit, the digital data sequence into a signal point sequence u(D) such that a component u_k of u(D) at a given time k is selected based in part on past components $\{y_{k-1}, y_{k-2}, \ldots\}$ of a channel output sequence y(D)=x(D)h(D) based on feedback information provided by a precoder,

a precoder for generating, using the plurality of registers and the arithmetic logic unit, said signal point sequence x(D) according to x(D)=u(D)+d(D), wherein d(D) represents a nonzero difference between a selected non-zero sequence c(D) and a postcursor intersymbol interference (ISI) sequence p(D) substantially of a form p(D)=x(D)[h(D)−1], wherein c(D) is selected such that the channel output sequence y(D) is a code sequence in said trellis code C,

and the receiving unit includes:

a decoding unit, operably coupled to the channel, for receiving and decoding a received sequence r(D) to provide an estimated output sequence $\hat{y}(D)$, and

a recovery unit, operably coupled to the decoding unit, for substantially recovering an estimate $\hat{u}(D)$ of the signal point sequence u(D).

40. The system of claim 39 wherein the decoding unit is a decoder for the trellis code C that receives and decodes a noisy received sequence r(D) which is of a form:

$$\begin{aligned} r(D) &= x(D)h(D) + w(D) \\ &= y(D) + w(D) \\ &= [u(D) + c(D)] + w(D), \end{aligned}$$

where w(D) represents additive white noise, to provide an estimate $\hat{y}(D)$ of the channel output sequence y(D)=x(D)h(D), and the recovery unit, operably coupled to the decoding unit, substantially recovers an estimate $\hat{u}(D)$ of the input sequence u(D).

41. The system of claim 39 wherein the components C_k of c(D) are selected from a time-zero lattice Λ_0 of said trellis code or a sublattice Λ_s thereof.

42. The device of claims 39 wherein said components u_k of u(D) are selected based on the state S_k of the channel output sequence y(D) in said trellis code.

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U.S. PATENT: *6,198,776*
ISSUE DATE: *March 06, 2001*

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Under Secretary of Commerce for Intellectual Property
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Certifying Officer

US006198776B1

(12) United States Patent
Eyuboglu et al.

(10) Patent No.: US 6,198,776 B1
(45) Date of Patent: *Mar. 6, 2001

(54) **DEVICE AND METHOD FOR PRECODING DATA SIGNALS FOR PCM TRANSMISSION**

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(73) Assignee: **Motorola Inc.**, Schaumburg, IL (US)

(*) Notice: Subject to any disclaimer, the term of this patent is extended or adjusted under 35 U.S.C. 154(b) by 0 days.

This patent is subject to a terminal disclaimer.

(21) Appl. No.: **08/999,249**

(22) Filed: **Dec. 29, 1997**

Related U.S. Application Data

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(51) **Int. Cl.**^{7} **H04L 25/49**
(52) **U.S. Cl.** **375/286**; 375/295
(58) **Field of Search** 375/286, 216, 375/242, 222, 295; 379/93.01

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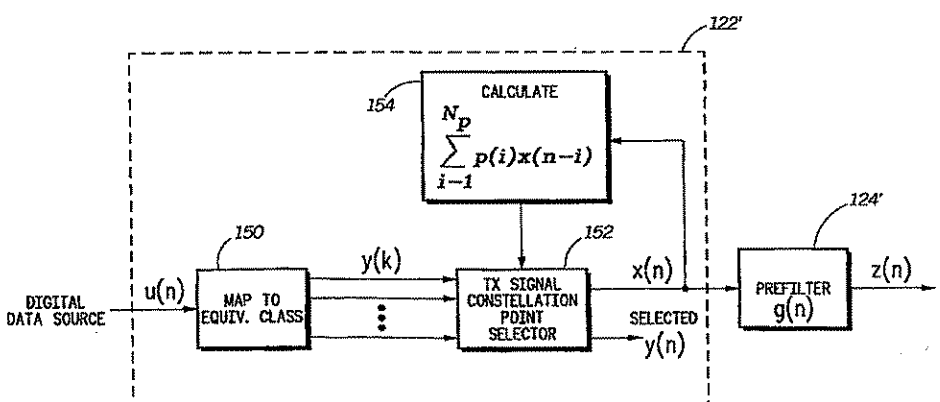
(57) **ABSTRACT**

A device and method for preceding data signals for pulse code modulation (PCM) transmission includes a transmitter for transmitting a sequence of analog levels over an analog channel to a quantization device, wherein the analog channel modifies the transmitted analog levels, the transmitter comprising: a mapping device for mapping data bits to be transmitted to a sequence of equivalence classes, wherein each equivalence class contains one or more constellation points; and a constellation point selector interconnected to the mapping device which selects a constellation point in each equivalence class to represent the data bits to be transmitted and which transmits an analog level that produces the selected constellation point at an input to the quantization device.

31 Claims, 10 Drawing Sheets



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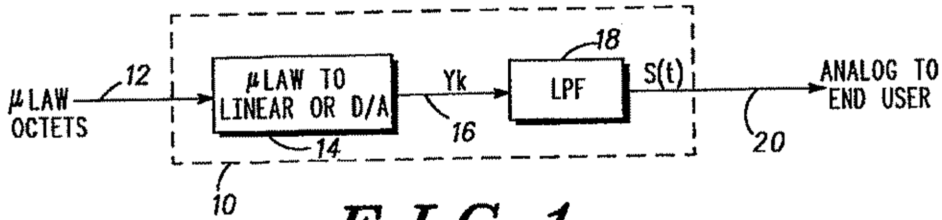


FIG.1

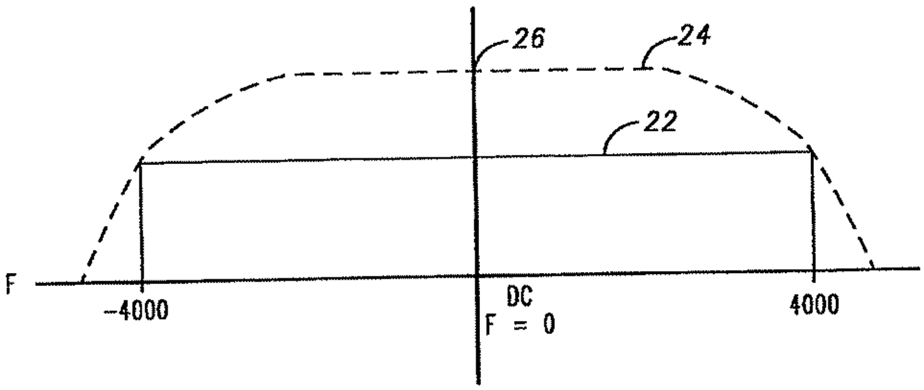


FIG.2

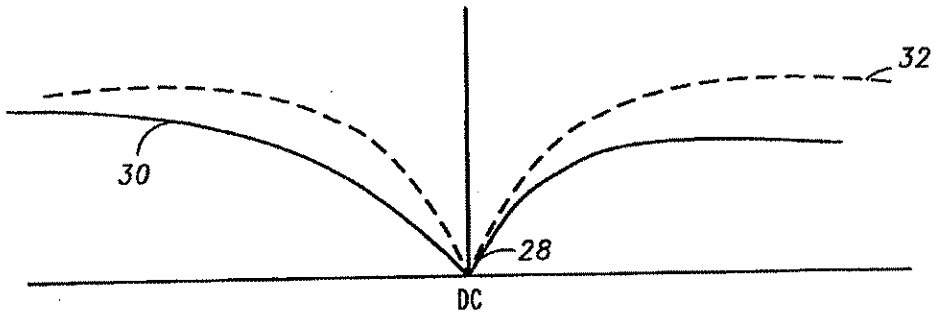


FIG.3

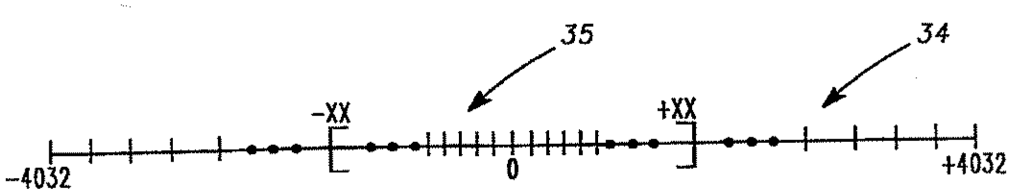


FIG.4

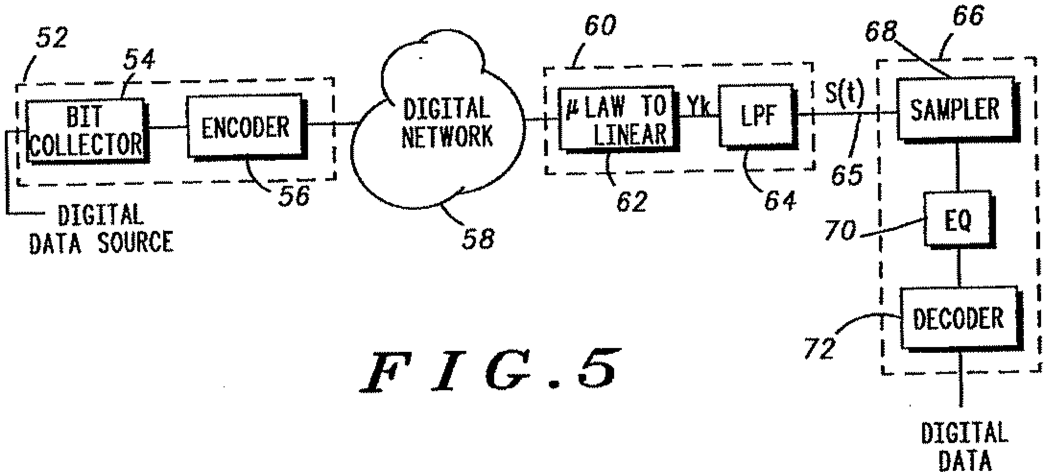


FIG. 5

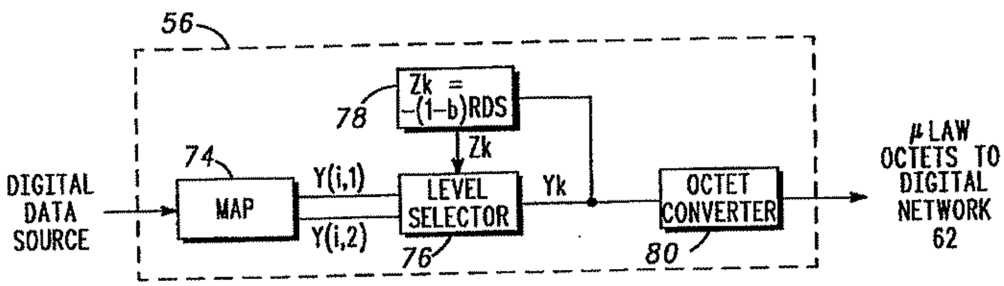


FIG. 6

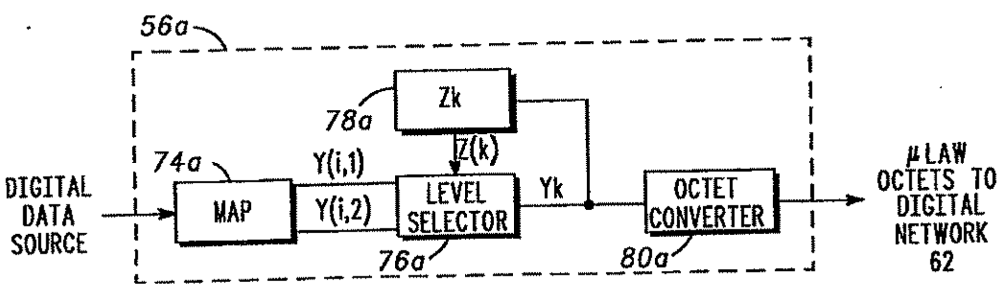
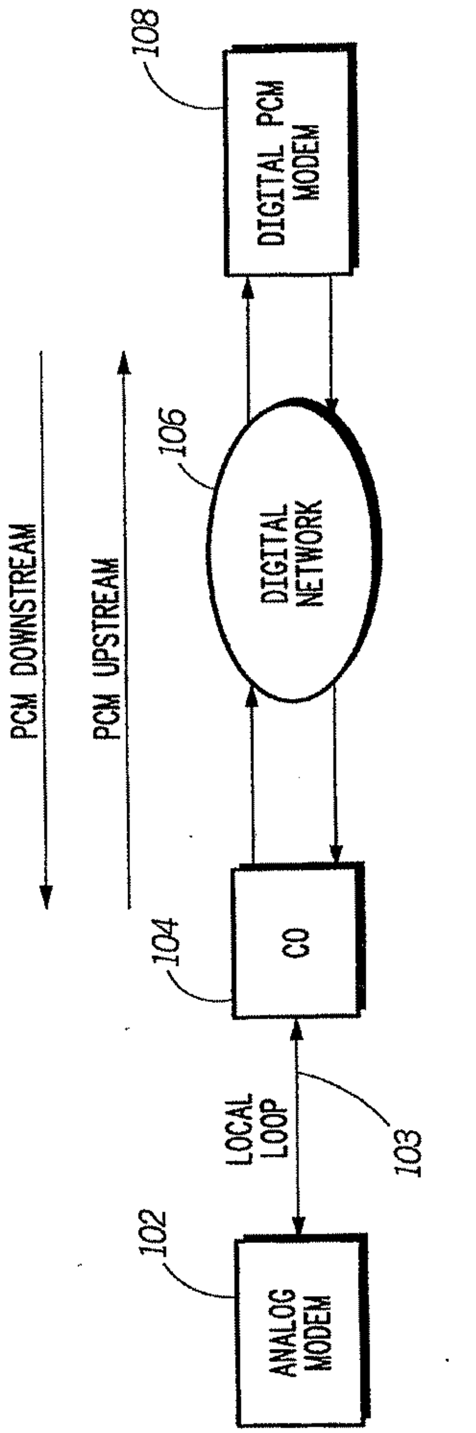


FIG. 7



100

PRIOR ART

FIG. 8

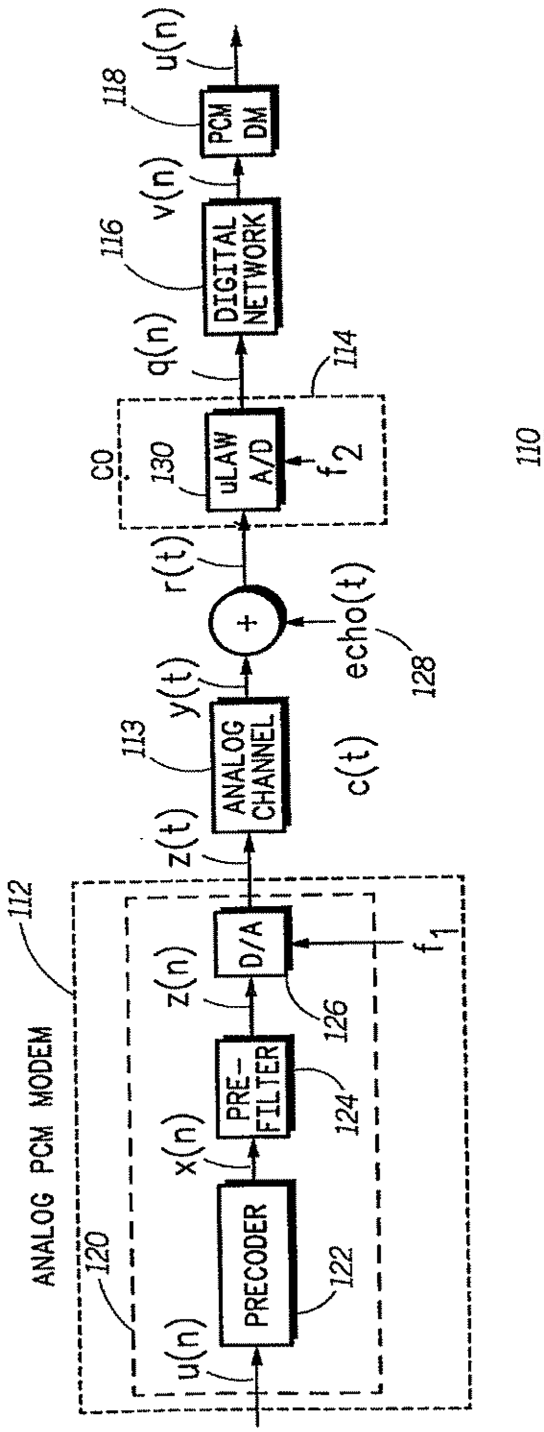
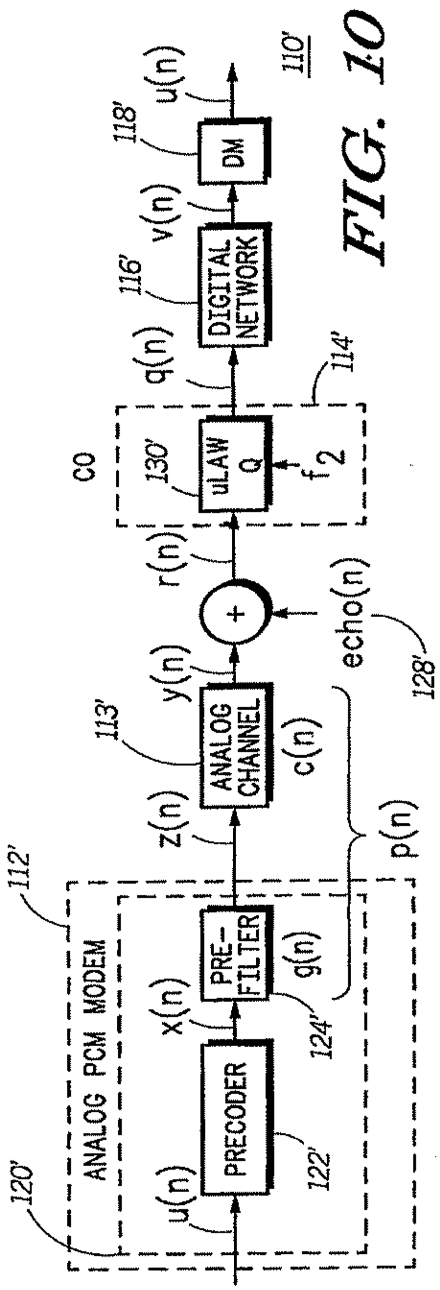
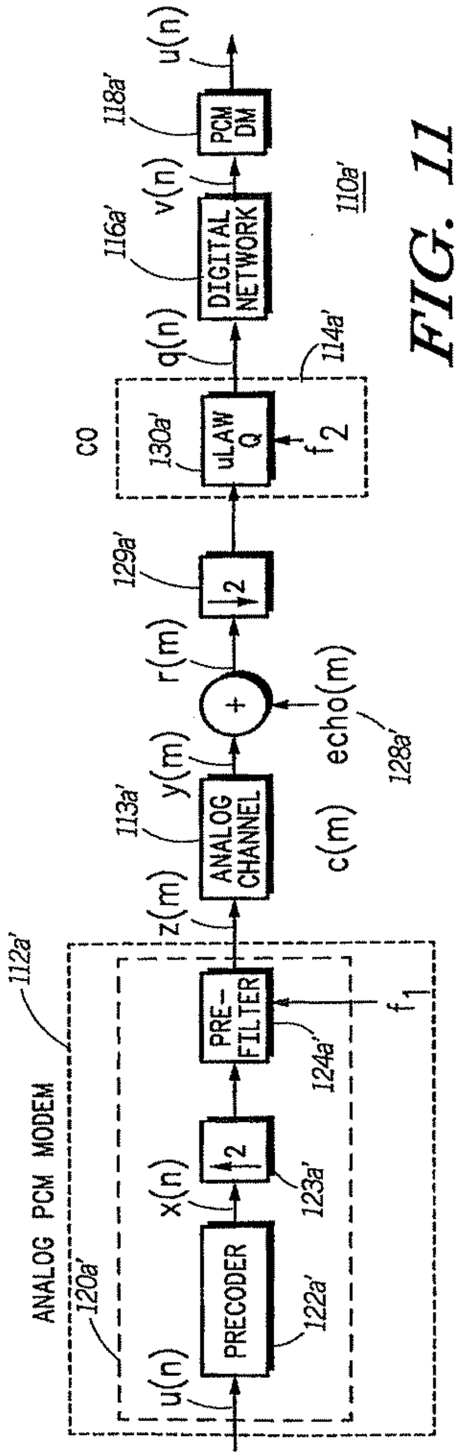

ANALOG PCM MODEM
112
120
u(n)
PRECODER
122
x(n)
PRE-FILTER
124
z(n)
D/A
126
f_1
z(t)
ANALOG CHANNEL
113
c(t)
y(t)
+
echo(t)
128
r(t)
CO
130
uLAW A/D
f_2
114
q(n)
DIGITAL NETWORK
116
v(n)
PCM DM
118
u(n)
110

FIG. 9



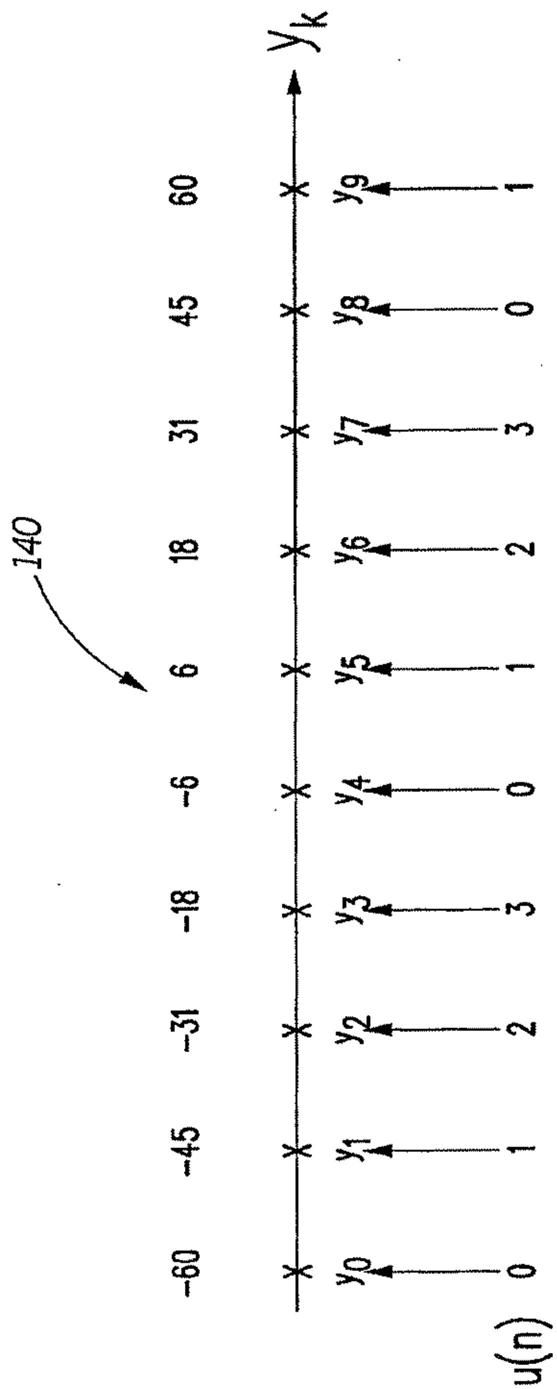
ANALOG PCM MODEM
120'
112'
u(n)
PRECODER
122'
x(n)
PRE-FILTER
g(n)
124'
z(n)
ANALOG CHANNEL
113'
c(n)
p(n)
y(n)
echo(n)
128'
r(n)
CO
uLAW Q
130'
f_2
114'
q(n)
DIGITAL NETWORK
116'
v(n)
DM
118'
u(n)
110'

FIG. 10



ANALOG PCM MODEM
112a'
120a'
u(n)
PRECODER
122a'
x(n)
↑2
123a'
PRE-FILTER
124a'
f_1
z(m)
ANALOG CHANNEL
113a'
c(m)
y(m)
echo(m)
128a'
r(m)
↓2
129a'
CO
uLAW Q
130a'
f_2
114a'
q(n)
DIGITAL NETWORK
116a'
v(n)
PCM DM
118a'
u(n)
110a'

FIG. 11



y_k
140
-60 -45 -31 -18 -6 6 18 31 45 60
y_0 y_1 y_2 y_3 y_4 y_5 y_6 y_7 y_8 y_9
0 1 2 3 0 1 2 3 0 1
u(n)

FIG. 12

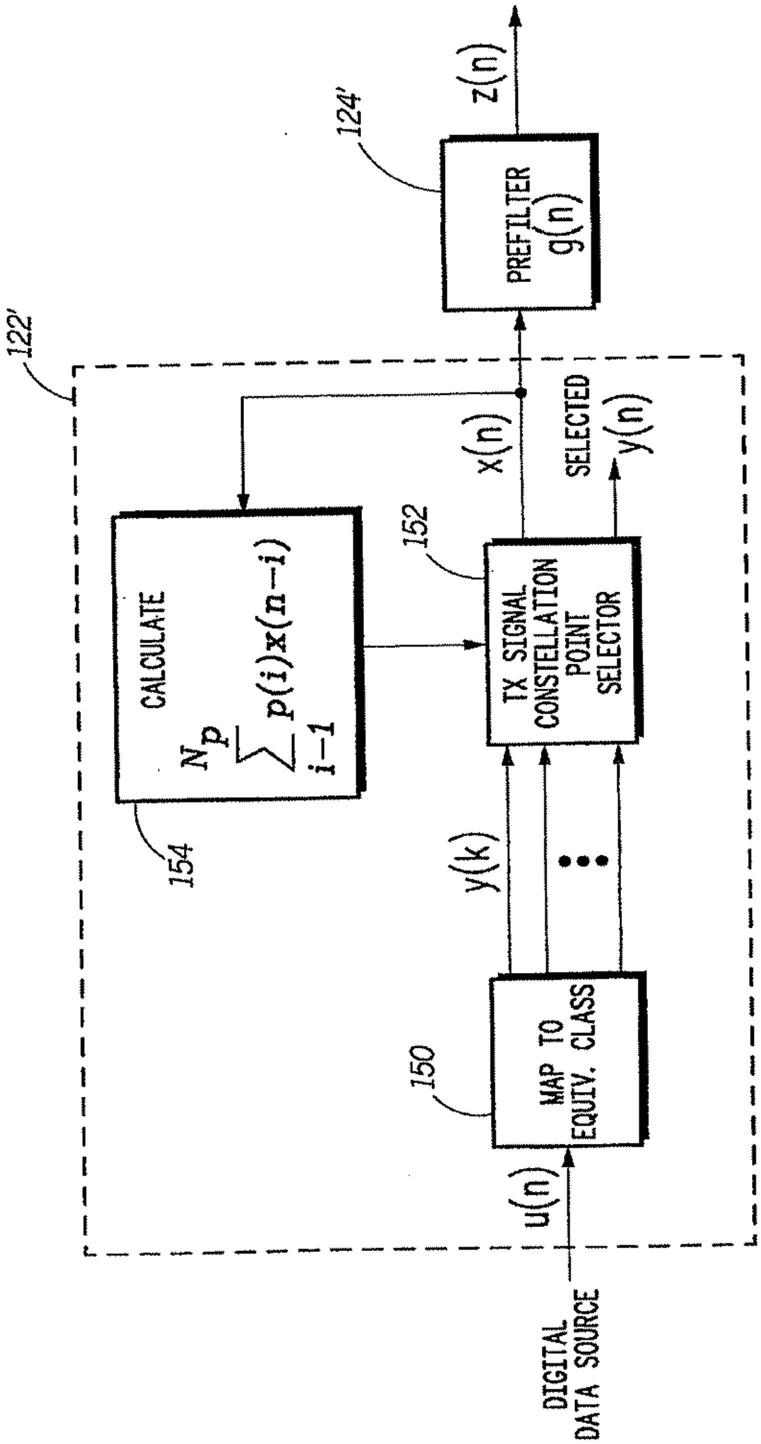

124'
z(n)
PREFILTER
g(n)
122'
CALCULATE
$\sum_{i-1}^{N_p} p(i)x(n-i)$
154
152
x(n)
SELECTED
y(n)
TX SIGNAL
CONSTELLATION
POINT
SELECTOR
y(k)
150
MAP TO
EQUIV. CLASS
u(n)
DIGITAL
DATA SOURCE

FIG. 13

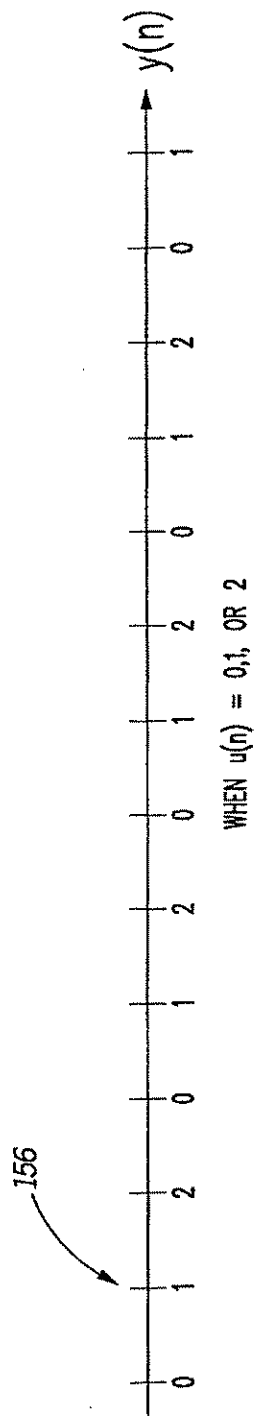


FIG. 14A

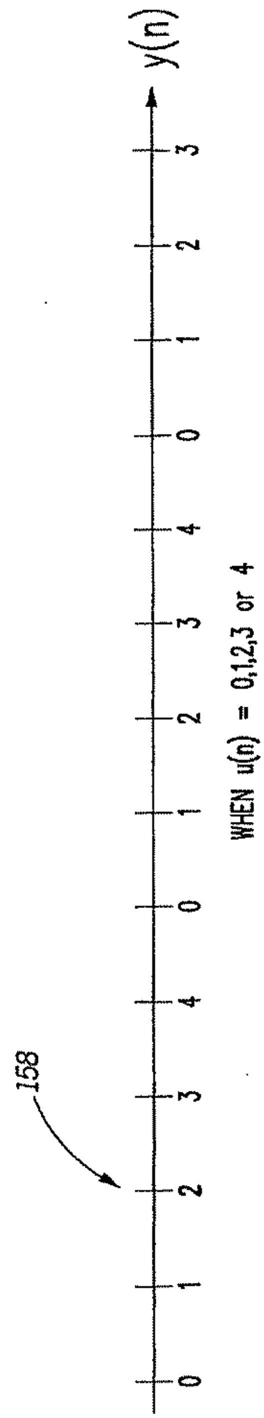
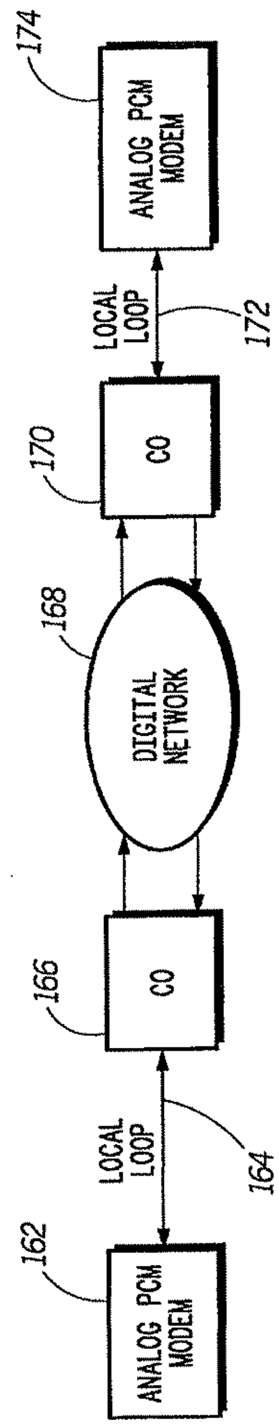


FIG. 14B



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PRIOR ART

FIG. 15

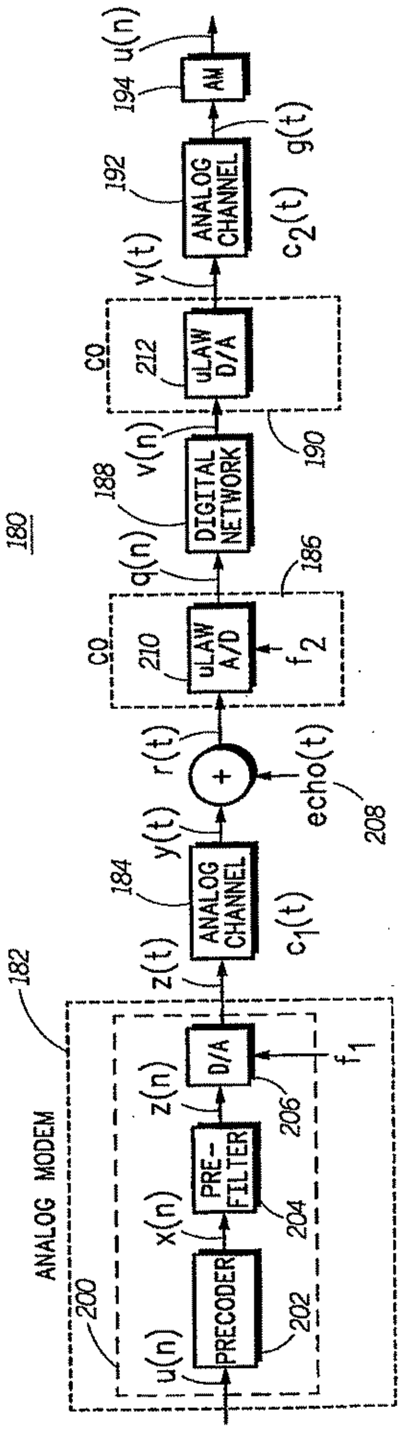
180
ANALOG MODEM
182
200
u(n)
PRECODER
202
x(n)
PRE-FILTER
204
z(n)
D/A
206
f1
z(t)
ANALOG CHANNEL
184
c1(t)
y(t)
+
echo(t)
208
r(t)
CO
uLAW A/D
210
f2
186
q(n)
DIGITAL NETWORK
188
v(n)
CO
uLAW D/A
212
190
v(t)
ANALOG CHANNEL
192
c2(t)
g(t)
AM
194
u(n)

FIG. 16

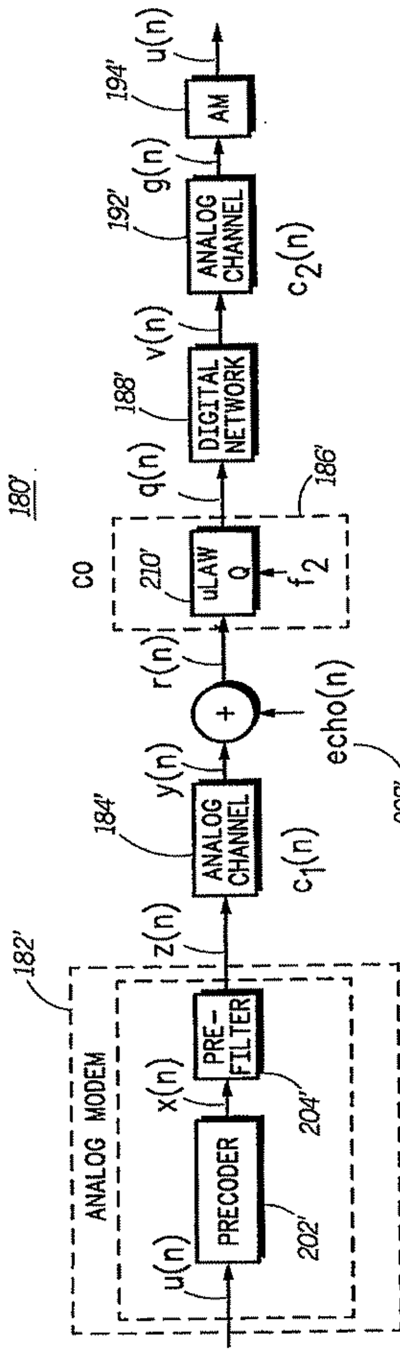
180'
ANALOG MODEM
182'
u(n)
PRECODER
202'
x(n)
PRE-FILTER
204'
z(n)
ANALOG CHANNEL
184'
c1(n)
y(n)
+
echo(n)
208'
r(n)
CO
uLAW Q
210'
f2
186'
q(n)
DIGITAL NETWORK
188'
v(n)
ANALOG CHANNEL
192'
c2(n)
g(n)
AM
194'
u(n)

FIG. 17

DEVICE AND METHOD FOR PRECODING DATA SIGNALS FOR PCM TRANSMISSION

RELATED APPLICATIONS

This application is a continuation-in-part of U.S. application Ser. No. 08/747,840, now U.S. Pat. No. 5,818,075 filed Nov. 13, 1996, which is hereby incorporated by reference in its entirety.

FIELD OF INVENTION

This invention relates to a device and method for precoding data signals for pulse code modulation (PCM) transmission.

BACKGROUND OF INVENTION

Conventional modems, such as V.34 modems, treat the public switched telephone network (PSTN) as a pure analog channel even though the signals are digitized throughout most of the network. In contrast, pulse code modulation (PCM) modems take advantage of the fact that most of the network is digital and that typically central site modems, such as those of internet service providers and on-line services, are connected to the PSTN via digital connections (e.g., T1 in the United States and E1 in Europe). First generation PCM modems transmit data in PCM mode downstream only (i.e., from a central site digital modem to an analog end user modem) and transmit in analog mode, e.g. V.34 mode, upstream (i.e., from the end user modem to the central site modem). Future generation PCM modems will also transmit data upstream in PCM mode.

With PCM downstream, the central site PCM modem transmits over a digital network eight bit digital words (octets) corresponding to different central office codec output levels. At the end user's central office, the octets are converted to analog levels which are transmitted over an analog loop. The end user's PCM modem then converts the analog levels, viewed as a pulse code amplitude modulated (PAM) signal, into equalized digital levels. The equalized digital levels are ideally mapped back into the originally transmitted octets and the data the octets represent.

With PCM upstream, the end user PCM modem transmits analog levels over the analog loop corresponding to data to be transmitted. The analog levels are modified by the channel characteristics of the analog loop and the modified levels are quantized to form octets by a codec in the end user's central office. The codec transmits the octets to the PCM central site modem over the digital network. The PCM central site modem determines from the octets the transmitted levels and from the levels the data transmitted by the end user PCM modem is recovered.

A difficulty that exists with upstream PCM transmission is that the levels transmitted by the end user PCM modem are modified by the analog loop. Since these modified levels are the levels that are quantized to form octets by the codec, and not the levels that are actually transmitted, it can be difficult for the central site modem to accurately determine from the octets the data being transmitted by the end user PCM modem. This difficulty is compounded by the fact that there is a channel null in the analog loop, quantization noise introduced by the codec in the end user's central office and downstream PCM echo, which make it more difficult for the central site PCM modem to accurately recover the data transmitted.

Therefore, a need exists for a device and method for precoding data signals for PCM transmission such that the analog levels that are transmitted by the end user PCM modem accurately produce predetermined analog levels (constellation points) at the input to the codec in the end user's central office, which analog levels (constellation points) correspond to the data to be transmitted by the end user PCM modem. Moreover, there is a need for a device, system and method for preceding data signals for PCM transmission which limits the transmit power and combats a channel null introduced by the analog loop and quantization noise introduced by the codec in the end user's central office.

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 is a simplified block diagram of a typical telephone company central office;

FIG. 2 is plot of the frequency spectrum of the y_k signals output from the μ-law to linear converter of FIG. 1 and the spectral shape of the low pass filter of FIG. 1;

FIG. 3 is a plot of a portion of two frequency spectrums each having a null at DC, wherein one spectrum falls off to zero very abruptly at DC and the other spectrum falls off more gradually;

FIG. 4 is a diagrammatic representation of a portion of a typical μ-law constellation;

FIG. 5 is a block diagram of a modem data connection over the telephone system including a transmitter for spectrally shaping signals according to this invention;

FIG. 6 is a block diagram of the encoder of FIG. 6 used specifically for creating a DC null in said analog signals over an analog loop of the PSTN;

FIG. 7 is a block diagram of the encoder of FIG. 6 which may be used generally for modifying, as desired, the frequency spectrum of the signals output from the analog loop to the end user;

FIG. 8 is a block diagram of a typical analog PCM modem to digital PCM modem communication system;

FIG. 9 is a more detailed block diagram depicting PCM upstream transmission according to this invention;

FIG. 10 is an equivalent discrete time block diagram of the block diagram of FIG. 9;

FIG. 11 is the equivalent discrete time block diagram of the block diagram of FIG. 9 with the analog modem sampling rate twice that of the CO sampling rate;

FIG. 12 is an example of a transmit constellation having equivalence classes according to this invention;

FIG. 13 is a more detailed block diagram of the analog PCM modem transmitter of FIG. 10 according to this invention;

FIG. 14A is another example of a transmit constellation having equivalence classes according to this invention;

FIG. 14B is yet another example of a transmit constellation having equivalence classes according to this invention;

FIG. 15 is a block diagram of a typical analog PCM modem to analog PCM modem communication system;

FIG. 16 is a more detailed block diagram depicting PCM transmission with the PCM modem communication system of FIG. 15; and

FIG. 17 is an equivalent discrete time block diagram of the block diagram of FIG. 16.

DETAILED DESCRIPTION OF A PREFERRED EMBODIMENT

There is first described below a technique for PCM downstream spectral shaping or preceding of data signals.

Then, there is described a preceding technique for PCM upstream transmission of data signals. Finally, it is described how the PCM upstream preceding technique according to this invention may be generalized for use in a PCM communication system interconnecting two analog PCM modems, as opposed to the typical analog PCM modem and digital PCM modem interconnection.

PCM Downstream Spectral Shaping/Precoding

FIGS. 1 and 2 illustrate the presence of energy near DC in the signals transmitted to a remote user's modem over an analog loop. There is shown in FIG. 1 a portion of a typical telephone central office on a PSTN which receives at input 12 μ-law octets transmitted from a modem (transmitting modem, not shown) directly attached to the digital portion of the telephone system, such as the one described in the co-pending applications referred to above which directly encodes the digital data into octets for transmission. These octets are converted by a D/A converter, also known as a μ-law to linear converter 14, to a sequence of voltage levels, y_k, each level being one of 255 μ-law levels. The levels are output over line 16 to a LPF 18 which outputs over analog loop towards the remote modem's receiver a filtered analog signal s(t) which is an analog representation of the levels. The analog signal is demodulated and decoded by the receiving modem which outputs a digital bitstream which is an estimate of the originally transmitted data.

The sequence of levels y_k on line 16 from μ-law to linear converter 14 has a flat frequency response 22, FIG. 2. The spectral shape 24 of LPF 18 contains a significant amount of energy near DC (f=0) as illustrated at point 26. Since the sequence y_k has a flat frequency response, the spectrum of the signal s(t) output by filter 18 has the same spectral shape 24 as the filter 18 and therefore the signal s(t) also contains a significant amount of energy near DC. As described above, this energy near DC tends to saturate the transformers on the system which produces unwanted non-linear distortion in the signal s(t) transmitted towards the receiving modem.

In some applications this distortion must be reduced. This can be accomplished by reducing the signal energy near DC in the transmitted signal. Such a DC null 28 is depicted in FIG. 3. As is known in the state-of-the-art, in order to create this spectral null at DC in the transmitted signal, the running digital sum (RDS) of the transmitted levels y_k (namely, the algebraic sum of all previously transmitted levels) must be kept close to zero. The shape of the spectrum around the DC null 28 can vary from a relatively shallow sloped spectrum 30 to a spectrum 32 which falls off very abruptly at DC. The sharpness of the null depends on how tightly the RDS is controlled.

The present invention accordingly encodes the digital data being transmitted into μ-law octets in a manner that maintains the RDS near zero to create the desired spectral null at DC thereby reducing the non-linear distortion caused by transformer saturation.

To illustrate the method of creating a spectral null, we consider an example of transmitting 6 bits with every symbol y_k. It will be apparent to those skilled in the art that the invention can be used for transmitting any other number of bits per symbol, or when the number of bits per symbol transmitted varies from symbol to symbol. In a system without a spectral null, one first selects a subset of 64 levels from the available 255 μ-law levels such that a minimum distance d_{min} between levels is maintained. These 64 levels are symmetric in the sense that for every positive level there is a negative level of the same magnitude. For example, one can achieve a d_{min} of 32 for an average energy well under −12 dBm0, the regulatory limit.

A partial representation of all 255 μ-law levels 34 (128 positive and 127 negative) is shown in FIG. 4. These levels follow a logarithmic law, with the 64 levels closest to the origin being uniformly spaced between −63 and 63 with a spacing of 2. The next positive and negative segments start at +/−66 and they each contain 16 points spaced by 4. The scale continues with segments of 16 points, each with a spacing of the form 2^n separated from the previous segment by a spacing of $0.75 * 2^n$. The final segments extend between +/−2112 and +/−4032 with a spacing of 128. The set 35 is the set of 64 levels selected from these 255 levels to represent each combination of six bits, i.e. $2^6=64$.

In the transmitter, incoming bits are collected in groups of 6, and then mapped into μ-law octets, which represent the desired level. In the central office, the μ-law octets are converted into levels, and the resulting levels are then transmitted. In the receiver, an equalizer compensates for the distortion introduced by the LPF and the local loop, and then a decision device estimates the transmitted level, by selecting the level that is closest to the received point.

In order to achieve spectral shaping in the above example, additional levels are also used, but the minimum distance between levels is still kept at 32. For example, consider the case where 92 levels are used. First, these 92 levels are divided into equivalence classes. There are a number of different ways for generating these equivalence classes. One particularly useful way is described here: we label the levels by integers 0 through 91, for example by assigning the label 0 to the smallest (most negative) level, the label 1 to the next smallest level, and so on. Then, we define 64 "equivalence classes" by grouping together levels whose labels differ exactly by 64. Such grouping leads to 36 equivalence classes with only one level corresponding to one of 36 innermost levels of smallest magnitude, and 28 equivalence classes with two levels whose labels differ by 64. Other methods for generating the equivalence classes may be used. Each possible combination of 6 bits to be transmitted is then represented by an equivalence class.

For example, the bit combination 000000 may correspond to the first equivalence class which consists of two levels each being represented by a different octet. Note that it is not necessary to use the full dynamic range of the D/A converter. The technique can work with any number of levels, as long as more than 64 levels are used. Of course, the more levels used, the better the desired spectral shape can be achieved. Our experiments indicate that very few additional levels need to be considered for generating a DC null with a relatively sharp notch.

In the above example, since each combination of six information bits is represented by an equivalence class and often there is more than one level in an equivalence class, the information bits must be mapped into one of the levels in a selected equivalence class before an octet representing that level is transmitted. This function is described below with regard to FIGS. 5–7.

Transmitter 52, FIG. 5, receives from a digital data source, such as a computer, a bitstream of digital data and with bit collector 54 divides the bits into groups of six, for example. Each six-bit group is provided to encoder 56 which selects the equivalence classes from which the desired levels to achieve the spectral null at DC will be selected. The octets which represent the selected levels are output from encoder 56, transmitted over digital circuit-switched telephone network 58 and arrive at the remote user's central office 60. At central office 60, the octets are converted by μ-law to linear converter 62 to the levels, y_k, which pass through LPF 64 and are output over local analog loop 65 as a signal s(t)

having a spectral null at DC. In receiver **66**, the signal s(t) is sampled by sampler **68**, an equalizer **70** compensates for the distortion introduced by LPF **64** and the local loop, and then a decision device or decoder **72** estimates the transmitted level by selecting the level that is closest to the received point. From the level the decoder **72** determines the equivalence class and then recovers the six information bits by performing an inverse mapping function.

The operation of receiver **66** is essentially unchanged as compared to the receiver described in the co-pending applications referred to above. The only difference is that the receiver now needs to consider a larger set of possible levels and the inverse mapping involves the determination of the equivalence class. Equalizer **70** compensates for the linear distortion introduced by the LPF **64** and the local loop **65**, as described in the co-pending applications. For example, when a linear equalizer is used, the output of the equalizer can be represented as follows:

$$rk=yk+nk \tag{1}$$

where nk is the total noise plus distortion present at the output of the equalizer. Decoder **72** then selects the levels y_k nearest to rk as the decision, determines its equivalence class, and then recovers the six information bits by an inverse map.

If the equalizer includes a maximum-likelihood sequence estimator (e.g., the Viterbi equalizer), then the received signal can be represented in the form

$$rk=\Sigma yk-j\ fj+nk, \tag{2}$$

and this time, the decoder selects the closest sequence $\{y_k\}$ using a Viterbi decoder. For each estimated symbol y_k, the decoder determines its equivalence class and then finds the six information bits via an inverse map.

Encoder **56**, FIG. **6**, includes MAP **74** which is a look-up table containing for each possible combination of the six-bit groups of data received from bit collector **54**, FIG. **5**, levels representing each equivalence class i, where i is an integer between **0** and **63**. Each level, two in this example, y(i,1) and y(i,2) is provided to level selector **76** where a decision is made as to which level, y_k, is to be transmitted.

This decision is made as follows. First, encoder **56** keeps track of the running digital sum (RDS) of the transmitted levels, y_k, by feeding back the output of level selector **76** to function block **78**. From the previously transmitted levels, y_k, function block **78** calculates the weighted RDS, $z_k=-(1-b)$RDS, where $0\leqq b<1$ is weighting factor. Because of D/A nonlinearities, the exact values of the y_k levels may not be known in encoder **56**; however, this should not have a significant effect. It is possible to determine the error and send this information back to encoder **56** to make these calculations more accurate.

Given the group of six bits to be transmitted, level selector **76** selects as the level y_k from the equivalence class {y(i,1), y(i,2)} the level closest to the weighted RDS. It can be seen that when the RDS is positive, zk will be negative and vice versa. This enables the encoder to choose a level, y_k, from each equivalence class such that when its value is added to the RDS it will bring it closer to zero than the other levels in the equivalence class. After selecting the level yk the octet which represents the level y_k is determined by octet converter **80** and transmitted over the digital network. The value of the transmitted octet can be obtained from a look-up table.

The variable b is a weighting factor that controls the trade-off between the sharpness of the spectral null and the average energy of the transmitted signal. Our analysis has shown that when the number of levels is sufficiently larger than the number of equivalence classes, the sequence y_k will have a spectrum which can be approximated by the filter response h(D)=(1-D)/(1-bD). Clearly, when b=0, we find that h(D)=1-D, which is the well-known Class I Partial Response with a sinusoidal spectral shape having a null at DC. On the other hand, as b approaches 1, the spectrum becomes flat across much of the band except for a very sharp spectral null at DC. It can be seen that for b=0, the average energy of y_k will be twice as large as in the case of a flat spectral shape. As b approaches 1, however, the average energy increase will disappear. In some applications, it may be desirable to keep the constellation expansion, measured by the ratio of the number of levels to the number of equivalence classes. It will be apparent to those skilled in the art that the invention can be used with constellations of any number of levels, and with any smaller number of equivalence classes.

The present invention may be more broadly utilized to spectrally shape, as desired, the analog signals output from the μ-law to linear converter at the central office. The example described above is a specific case of using this invention to reduce the energy of the transmitted signal around DC, but the principals of this invention used in that example can be generalized to spectrally shape signals in numerous ways, for example, to pre-equalize the signals.

A generic version of the encoder of this invention, encoder **56a**, is shown in FIG. 7. The only difference between this general case and the special case of a spectral null described above is how the sequence or spectral function zk is generated. Let h(D) be a monic, causal impulse response of a filter representing the desired spectral shape, where D is a delay operator. Suppose we represent the sequences {yk} and {zk} using D-transform notation as y(D) and z(D), respectively. Then, the sequence z(D) can be represented as

$$z(D)=(1-1/h(D))y(D). \tag{3}$$

A close examination of this equation reveals that at a given time k, zk only depends on past values of yk, and therefore can be determined recursively. Thus, for each six bit group, encoder *56a* determines which level from the associated equivalence class is nearest in value to z_k and selects that level. The octet representing that level is then transmitted. Again, our analysis shows that for sufficiently large number of levels the sequence $\{y_k\}$ transmitted by the central office **60** will have a spectrum closely approximating the spectrum of the filter with response h(D).

The technique described here can also be used in conjunction with a more complex scheme for mapping the information bits to equivalence classes. For example, it can be used in conjunction with shell mapping, a mapping technique used in the V.34 high-speed modem specification.

The examples described above are for an uncoded system. However, the principals can be easily applied to a coded system, for example a trellis coded system. The only difference in this case is that the equivalence classes are further partitioned into subsets, which are used to construct the trellis code.

For example, when a one-dimensional trellis code based on a 4-way set partition is utilized together with the same 64-level signal constellation to send bits per symbol, the equivalence classes are partitioned into subsets as follows: $a_1, b_1, c_1, d_1, a_2, b_2, c_2, d_2, \ldots a_n, b_n, c_n, d_n$. In the example described above, the 64 equivalence classes would be partitioned into four subsets each containing sixteen equivalence classes. The output of a rate-½ convolutional encoder,

e.g. two of the six bits in a group, then determines the subset, and the remaining four "uncoded" bits select the specific equivalence class within the subset. The actual level from the chosen equivalence class in the chosen subset is selected as described above. The operation of the encoder is otherwise unchanged.

Of course, when trellis coding is utilized, the receiver will use a decoder to select the most likely sequence. The trellis decoder may also be an equalizer, jointly decoding the trellis code and equalizing for intersymbol interference.

It may also be possible to use the present invention to enable detection of loss of frame synchronization in a receiver. This can be accomplished by infrequently, but periodically violating the rule for selecting the signal point in a given equivalence class, where the period is chosen to be an integer multiple of the desired framing. A loss of frame synchronization, can be detected in the receiver by monitoring such rule violations. The receiver can also reacquire frame synchronization or may simply request a synchronization pattern (training sequence) from the transmitter.

PCM Upstream Precoding

There is shown in FIG. 8, a typical PCM communication system **100**. System **100** includes analog PCM modem **102** connected to a telephone company central office (CO) **104** over a local analog loop or channel **103**. There is also included a digital network **106** which is interconnected to CO **104** and to digital PCM modem **108**. With this system, PCM data may be transmitted both in the downstream direction (i.e., from digital PCM modem **108** to analog PCM modem **102**) and in the upstream direction (i.e., from analog PCM modem **102** to digital PCM modem **108**). This type of bidirectional PCM communication system is described in U.S. application Ser. No. 08/724,491, entitled Hybrid Digital/Analog Communication Device, which is assigned to the assignee of the present invention and which is incorporated herein in its entirety by reference.

In the above section a technique for PCM downstream spectral shaping or precoding of data signals is described. In this section there is described a precoding technique for PCM upstream precoding of data signals.

In FIG. 9 there is shown in block diagram **110**, an example of PCM upstream transmission in accordance with this invention. In block diagram **110** there is included analog PCM modem **112** interconnected to analog channel **113**. Analog PCM modem **112** includes transmitter **120** having a precoder **122**, prefilter **124** and a digital to analog converter (D/A) **126**. Precoder **122** receives digital data u(n) and outputs precoded digital data signal x(n). The precoded digital data signal is filtered by prefilter **124** to form signal z(n) which is provided to D/A **126**. D/A **126** converts the filtered signal z(n) to analog form and transmits analog signal, z(t), over analog channel **113**, having a channel characteristic, c(t).

The analog channel modifies the transmitted signal z(t) to form signal y(t). The signal y(t) then encounters downstream PCM echo, echo(t) **128**, that is added to y(t), producing signal r(t). Signal r(t) is received by μ-law (A-law in some countries outside of the US) quantizer **130** in central office (CO) **114** and is quantized according to the μ-law. See International Telecommunications Union, Recommendation G.711, Pulse Code Modulation (PCM) of Voice Frequencies, 1972.

The quantized octets (digital values), q(n), are transmitted over digital network **116** at a frequency of 8 kHz where they may be affected by various digital impairments, as discussed below. The possibly affected octets, v(n), are received by digital PCM modem **118** which ideally decodes the octets, v(n), into their corresponding constellation points, y(t), from which the original digital data, u(n), can be recovered. The decoding of v(n) is described in co-pending application Ser. No. 08/999,254 entitled System, Device and Method for PCM Upstream Transmission Utilizing an Optimized Transmit Constellation, CX097028, which is assigned to the assignee of the present invention and which is incorporated herein in its entirety by reference.

Before data can be transmitted upstream, the clock (f_1) of D/A **126** in analog PCM modem **112** must be synchronized to the clock (f_2) of CO **114**. This can be achieved by learning the clock from the downstream PCM signal (not shown) and synchronizing the clocks using the technique proposed in U.S. Pat. No. 5,199,046, entitled First and Second Digital Rate Converter Synchronization Device and Method, incorporated herein by reference in its entirety. Once the clocks are synchronized, PCM upstream block diagram **110**, FIG. 9, can be represented as equivalent discrete time block diagram **110'**, FIG. 10, with like components being represented by the same reference numbers containing a prime ('). In block diagram **110'** it is assumed that $f_1=f_2$; however, it must be noted that f_1 does not have to be equal to f_2 as long as the two clocks are synchronized. When f_1 is equal to f_2, n is the time index for 8 kHz samples, since the clock (f_2) of CO 24 is fixed at that frequency.

An example where f_1 does not equal f_2 is depicted in FIG. 11. Equivalent discrete time block diagram **110a'**, FIG. 11, is the same as equivalent discrete time block diagram **110'**, FIG. 10, except that there is a 2× up-sampler **123a'** in transmitter **120a'** and a 2× down-sampler **129a'** to account for the fact that $f_1=2f_2$. The variables "m" and "n" are the time indexes for 16 kHz and 8 kHz samples, respectively.

Precoder **122'** and prefilter **124'**, according to this invention, are designed to transmit signal z(n) over analog channel **113** such that predetermined constellation points, y(n), corresponding to digital data u(n) are produced at the input of μ-law quantizer **130'** (in combination with an echo component, echo(n), if present). In other words, the input of μ-law quantizer **130'** is y(n)+e(n) in the presence of echo(n) and just y(n) in the absence of echo(n).

Using the PCM upstream precoding technique described below, or another precoding technique, it is difficult for digital PCM modem **118'** to accurately decode u(n) from v(n) in the presence of echo, quantization and digital impairments without a properly designed transmit constellation of points, y(n). It is described in co-pending application CX097028 how to design the transmit constellation for y(n) to enable y(n) (and eventually u(n)) from v(n)) to be decoded in the presence of echo, quantization and digital impairments with minimized error probability.

As described in co-pending application CX097028, for a given connection, depending on the line conditions, a transmit constellation for each robbed bit signaling (RBS) time slot is selected. As an example, transmit constellation **140** is depicted in FIG. 12. This constellation includes ten constellation points, y_0–y_9, ranging in value from −39 to 39. It should be noted that the constellation points, y(n), are not necessarily G.711 μ-law levels.

The constellation points y(n) correspond to digital data to be transmitted, u(n). In other words, each constellation point represents a group of data bits and the number of data bits represented by each constellation point depends on the number of points in the constellation (and the number of equivalence classes which are described below). The more points in the constellation, the more bits of data that can be represented. As shown in FIG. 12, digital data u(n) is divided into four groups of bits **0**,**1**,**2** and **3**, corresponding to 00, 01,

10 and 11, for example. Thus, in this example each constellation point transmitted represents two bits and since the constellation points are transmitted at 8 k/sec, the data rate is 16 kbps. It must be understood that this is a simplified example and data may be mapped into u(n) using any mapping schemes that can map bits into equivalence classes, such as shell mapping or modulus conversion.

According to this invention, the constellation points are grouped into equivalence classes. An equivalence class is a set of typically two or more constellation points which represent the same group of bits or digital data to be transmitted, u(n). With constellation **140**, it is shown that constellation points $y_0(-60)$, $y_4(-6)$, and $y_8(45)$ form the equivalence class for u(n)=0. Constellation points $y_1(-45)$, $y_5(6)$, and $y_9(60)$ form the equivalence class for u(n)=1 and constellation points $y_2(-31)$, and $y_6(18)$ form the equivalence class for u(n)=2. Finally, constellation points $y_3(-18)$, and $y_7(31)$ form the equivalence class for u(n)=3.

Equivalence class selection is generally accomplished as follows. The constellation, with M points, is indexed as y_0, $y_1, \ldots y_{M-1}$ in ascending (or descending) order. Assuming u(n) has U values, e.g. U=4 as in the above example, then the equivalence class for u(n)=u contains all the y_k's where k modulo U is u. For example, in FIG. **11**, the equivalence class for u(n)=0 is y_0, y_u, y_{2U}, where U=4. Note that each equivalence class is not required to have the same number of constellation points.

The number of supporting data levels for u(n) should be chosen to satisfy the following two conditions: 1) The expansion ratio, which is defined as the ratio between the number of constellation points for y(n) and the number of supporting data levels for u(n), i.e., M/U; and2) TX power constraints.

The expansion ratio should be large enough to guarantee stable operation. The size of the expansion ratio will depends on the channel characteristics. In voice band modem applications, there is at least one spectral null at f=0. Therefore, we should have an expansion ratio of M/U≧2 to make the system stable. In practice, to guarantee the stability, the quality of the channel is determined from the channel response, c(n), and the minimum expansion ratio is set accordingly. For example, we can use C(f=4 kHz), the frequency response of the channel at 4 kHz (with respect to other frequencies like 2 kHz), as the quality of the channel and depending on that quality we set the minimum expansion ratio. If the C(f=4 kHz) C(f=2 kHz), then we set M/U≧2.0. As the C(f=4 kHz) gets smaller and smaller, the expansion ratio must be increased.

As described below, precoder **122'** selects the appropriate constellation point, y_k, from the equivalence class for the data, u(n), to be transmitted and determines a value for x(n) that will produce the selected constellation point at the input to μ-law quantizer **130'**.

The preceding scheme, i.e., the design of precoder **122'** and prefilter **124'**, are now described as follows. From the characteristics of analog channel **113'**, c(n), n=0,1, . . . N_c-1, determined by digital PCM modem **118'**, as described in co-pending application Ser. No. 08/999,416 entitled Device and Method for Detecting PCM Upstream Digital Impairments in a Communication Network, CX097029, which is assigned to the assignee of the present invention and which is incorporated herein in its entirety by reference, an optimal target response p(n), n=0, 1, . . . N_p-1, and corresponding prefilter g(n), n=−Δ,−Δ+1, . . . , $-\Delta+N_g-1$ (where Δ is the decision delay), as shown in FIG. **10**, are determined. This problem is similar to determining the optimal feedforward and feedback filters for a decision feedback equalizer (DFE). The prefilter corresponds to feedforward filter of DFE and the target response corresponds to feedback filter of DFE. See, N. Al-Dhahir, et al, "Efficient Computation of the Delay Optimized Finite Length MMSE-DFE", IEEE Transactions On Signal Processing, vol. 44, no. May 5, 1996, pp.1288–1292. Preferably, the target response p(n) and the filter g(n) will be determined in the analog modem, but they can be determined in the digital modem and transmitted to the analog modem.

The prefilter g(n), n=−Δ,−Δ+1, . . . , $-\Delta+N_g-1$, and the target response p(n), n=0, 1, . . . , N_p-1, (where p(0)=1) can be derived given c(n) by minimizing the cost function ζ as follows:

$$\zeta=\|g(n)*c(n)-p(n)\|^2+\alpha\|g(n)\|^2 \quad (4)$$

The first term ensures small intersymbol Interference (ISI), i.e., the receiver of digital PCM modem **118'** receives what precoder **122'** tried to encode, and the second term enforces the transmit (TX) power to stay finite and small. The term α is a constant term which should be chosen depending on the application. The larger α is the lower the TX power will be, but at the expense of ISI. A smaller α will give less ISI at the expense of TX power. Therefore α should be chosen depending on what is desired for ISI and TX power for a given application. As an example, α can be chosen to be the signal to noise ratio (SNR) of the system, which is $\sigma_n^2/E(x^2)$ or SNR normalized by channel energy, i.e., $SNR/\|c\|^2$. For $E(x^2)$, we can use −9 dBm which is the power constraint for upstream transmission. This minimization problem is the same as DFE tab initialization problem. The term σ_n^2 can be determined as described in co-pending application CX097028.

The initially determined p(n) and g(n) can always be used if the analog channel c(n) is time invariant. However, in practice, c(n) is time variant, though it is very slowly changing. Therefore, some kind of adaptation scheme is necessary. One way to do it is to monitor performance and retrain if the performance goes bad, i.e., re-estimating c(n) in the digital modem **118'** and sending a new c(n) back to analog modem **112'** to recalculate g(n) and p(n). Another way is to feedback the analog channel error signal, error(n), as described in co-pending application CX097029, from digital modem **118'** to analog modem **112'** through downstream data transmission and use that error signal to adapt p(n) and g(n).

Once the target response p(n) is determined precoder **122'** can be implemented. As explained above, we can send data u(n) by transmitting x(n) such as to produce at the input to quantizer **130'**, FIG. **10**, a constellation point y(n) which is one of the points in the equivalence class of u(n). Which constellation point from the equivalence class of u(n) to use to represent u(n) is usually selected to minimize the TX power of transmitter **120'**. The TX power of transmitter **120'** is the power of z(n) (or some other metric). In practice, since it is hard to minimize the power of z(n), the power of x(n) is minimized instead, which is a close approximation of minimizing z(n).

The following is a known relationship among x(n), y(n) and p(n):

$$y(n)=p(n)*x(n) \quad (5)$$

where "*" represents convolution. That relationship can be expressed as follows:

$$y(n)=p(0)x(n)+p(1)x(n-1)+\ldots p(N_p)x(n-N_p) \quad (6)$$

Since p(0) is designed to equal to 1, then equation (6) can be simplified as follows:

$$x(n) = y(n) - \sum_{i=1}^{N_p} p(i)x(n-i). \tag{7}$$

And, since p(n) and the past values of x(n) are known, the appropriate y(n), among the constellation points of the equivalence class of a given u(n), may be selected to minimize $x^2(n)$ in order to minimize the TX power of transmitter 120'.

Or, lookahead (i.e., decision delay) can be introduced to choose y(n). That is, y(n−Δ) can be chosen from the set of equivalence classes for u(n−Δ) to minimize

$$|x(n-\Delta)|^2+|x(n-\Delta+1)|^2+ \ldots +|x(n)|^2,$$

where:

$$x(n-j) = y(n-j) - \sum_{i=1}^{N_p} p(i)x(n-j-i) \tag{8}$$

where j=0, 1, . . . Δ and where y(n−j) is chosen from the set of equivalence classes of u(n−j) (j=0, 1, . . . Δ−1).

Precoder 122' may be implemented according to this invention as depicted in FIG. 13. Precoder 122' includes a mapping device 150 which receives the incoming digital data u(n) from a digital data source and, depending on the number of bits that can be transmitted with each constellation point, determines for each group of bits the equivalence class associated with the group of bits. Mapping device 150 outputs the constellation points, y_k, forming the equivalence class to TX signal/constellation point selector 152 which selects the constellation point, y_k, from the equivalence class and determines the transmit signal x(n) based on the input from calculation device 154.

Filter device 154 receives the transmit signal x(n) and calculates the summation term (or running filter sum (RFS)) of equation (7) above. Based on the value of the RFS, TX signal/constellation point selector 152 selects the constellation point in the equivalence class that will cause x(n) in equation (7) to be closest in value to zero and calculates the value of x(n) from the calculated RFS and the selected constellation point. The calculated transmit signal x(n) is then provided to prefilter 124' where x(n) is filtered to form signal z(n) which is transmitted over analog channel 113', FIG. 10.

In order to limit the TX power of transmitter 120', FIG. 10, to keep it within the FCC regulations, the equivalence classes for u(n) must be designed accordingly. With a constellation having a predetermined number of constellation points, If we want to send more data, then more groups of data, u(n), and hence equivalence classes for u(n) will be required. As a result, the constellation points will be further away and will require more transmit power. This is because y(n) is chosen as described below according to equation (7) to minimize $x^2(n)$. Therefore, if the constellation points in the equivalence classes are spaced further apart, it is more likely that $x^2(n)$ will be larger. Thus, to reduce the TX power, we can make the equivalence class of u(n) closer at the expense of rate. This is depicted in FIGS. 14A and 14B.

In FIGS. 14A and 14B, both constellations 156, FIG. 14A, and 158, FIG. 14B, have the same number of constellation points; however, constellation 156 has only three equivalence classes u(n)=0,1 and 2 while constellation 158 has five equivalence classes u(n)=0,1,2,3 and 4. Using constellation 158 will require more TX power than constellation 156, but it will be capable of transmitting at a higher data rate.

The approximate TX power (the power of z(n)) can be calculated as follows when U is the number of points desired to support u(n):

$$P_t \approx |g(n)|^2 \frac{1}{U} \sum_{i=0}^{U-1} dist^2(u(n)=i)/12 \tag{9}$$

where $|g(n)|^2$ is the energy of prefilter and dist(u(n)=i) is the minimum distance between the points in the equivalence points. For example, in FIG. 12 dist(u(n)=0)=|−6−(−60)|=54. Several values of U should be tried to find out the one which satisfies the power constraints. Note also that this should be done for each time slot.

The transmit constellation selection and equivalence class selection according to this invention may be summarized as follows:

1) Obtain digital impairments, calculate noise variance, σ_n^2, and echo variance, σ_e^2, as described in co-pending application CX097028;
2) From σ_e^2, σ_n^2, and the digital impairments, choose the proper constellation for y(n) for each time slot, also as described in co-pending application CX097028; and
3) For each time slot, find the number of points that can be supported for u(n) while satisfying the TX power constraints and the minimum expansion ratio to guarantee stable operation. From this U the constellation for y(n), and the equivalence classes for u(n) can be determined.

The above preceding technique which utilizes a one dimensional constellation can be expanded to multi-dimensional constellations by expanding the definition of the equivalence class of u(n). The following references describe various downstream preceding techniques using multi-dimensional constellations: Eyuboglu, Vedat; "Generalized Spectral Shaping for PCM Modems," Telecommunications Industry Association, TR30.1 Meeting, Norcross, Ga., Apr. 9–11, 1997, pages 1–5; Eyuboglu, Vedat; "Convolutional Spectral Shaping," Telecommunications Industry Association, Tr30.1 Meeting, Norcross, Ga., Apr. 9–11, 1997; Eyuboglu, Vedat; "More on Convolutional Spectral Shaping," ITU Telecommunications Standardization Sector 009, V.pcm Rapporteur Meeting, La Jolla, Calif., May 5–7, 1997; Eyuboglu, Vedat; "Draft Text for Convolutional Spectral Shaping," ITU-T SG 16 Q23 Rapporteur's Meeting, Sep. 2–11, 1997, Sun River, Oreg.; Eyuboglu, Vedat; "A Comparison of CSS and Maximum Inversion," Telecommunications Industry Association, Tr30.1 Meeting on PCM Modems, Galveston, Tex., Oct. 14–16, 1997; and Eyuboglu, Vedat; "Draft Text for Convolutional Spectral Shaping," Telecommunications Industry Association, Tr30.1 Meeting Galveston, Tex., Oct. 14–16, 1997.

Moreover, the example described above is for an uncoded system. However, the principals can be easily applied to a coded system, for example a trellis coded system. The only difference in this case is that the equivalence classes are further partitioned into subsets, which are used to construct the trellis code.

Generalized PCM Precoding

The above described PCM upstream precoding technique (i.e. from analog PCM modem 112', FIG. 10, to digital PCM modem 118, may be applied to an analog PCM modem to analog PCM modem connection as depicted in FIG. 15. System 160 includes analog PCM modem 162 connected to CO 166 over analog loop or channel 164. CO 166 is interconnected to digital network 168. Similarly analog PCM modem 174 is connected to CO 170 over analog loop or channel 172. And, CO 170 is connected to digital network 168.

Block diagram 180, FIG. 16, depicts an analog PCM modem to analog PCM modem connection according to this invention. In block diagram 180 there is included analog PCM modem 182 interconnected to analog channel 184. Analog PCM modem 182 includes transmitter 200 having a precoder 202, prefilter 204 and a digital to analog converter (D/A) 206. Precoder 202 receives digital data u(n) and outputs precoded digital data x(n). The precoded digital data is filtered by prefilter 204 to form signal z(n) which is provided to D/A 206. D/A 206 converts the filtered signal z(n) to analog form and transmits analog signal, z(t), over analog channel 184, having a channel characteristic, c(t).

The analog channel modifies the transmitted signal z(t) to form signal y(t). The signal y(t) then encounters PCM echo, echo(t) 208, that is added to y(t), producing signal r(t). Signal r(t) is received by μ-law (A-law in some countries outside of the US) quantizer 210 in central office (CO) 186 and is quantized according to the μ-law. See International Telecommunications Union, Recommendation G.711, Pulse Code Modulation (PCM) of Voice Frequencies, 1972.

The quantized octets (digital values), q(n), are transmitted over digital network 188 at a frequency of 8 kHz where they may be affected by various digital impairments, as discussed below. The possibly affected octets, v(n), are received by CO 190 and the octets, v(n), are converted by μ-law D/A 212 into analog levels for transmission over analog channel 192. The levels are received by analog PCM modem 194 which converts the levels to data u(n).

Once the clocks f1 to f2 of D/A 206 and D/A 210 are synchronized, block diagram 180 can be modeled as discrete time block diagram 180', FIG. 17. Analog PCM modem should do the equalization to get v(n) from g(n) in the same way as a downstream PCM modem works as is known in the art. Then, from v(n), a PCM upstream decoding algorithm to decode y(n), i.e. u(n), is undertaken.

The above only describes transmission from analog PCM modem 182' to analog PCM 194'; however, transmission in the other direction is accomplished in the same manner. The above described PCM upstream preceding technique (i.e. from analog PCM modem 112', FIG. 10, to digital PCM modem 118,) can be applied directly to an analog PCM modem to analog PCM modem connection as depicted in FIGS. 15–17.

It should be noted that this invention may be embodied in software and/or firmware which may be stored on a computer useable medium, such as a computer disk or memory chip. The invention may also take the form of a computer data signal embodied in a carrier wave, such as when the invention is embodied in software/firmware which is electrically transmitted, for example, over the Internet.

The present invention may be embodied in other specific forms without departing from the spirit or essential characteristics. The described embodiments are to be considered in all respects only as illustrative and not restrictive. The scope of the invention is, therefore, indicated by the appended claims rather than by the foregoing description. All changes which come within the meaning and range within the equivalency of the claims are to be embraced within their scope.

What is claimed is:

1. A transmitter which defines an equivalence class at the input of a quantization device for precoding a sequence of analog levels to be transmitted over an analog channel to said quantization device, the precoded sequence forming the input to the quantization device, comprising:

a precoder including a mapping device for mapping data bits to be transmitted to as a sequence of equivalent classes, wherein each equivalence class contains one or more constellation points; and a constellation point selector interconnected to the mapping device which selects a constellation point in each equivalence class to represent the data bits to be transmitted and which transmits a level that produces the selected constellation point to an input of the quantization device.

2. The transmitter of claim 1 further including a filter device, operably coupled to the constellation point selector, which receives at its input previously transmitted levels and provides its output to the constellation point selector.

3. The transmitter of claim 2 wherein the constellation point selector selects the constellation point from each equivalence class based on the output of the filter device.

4. The transmitter of claim 3 further including a prefilter, having a predefined filter response, g(n), for filtering the level transmitted by the constellation point selector.

5. The transmitter of claim 4 wherein the response of the filter device is:

$$\sum_{i=1}^{N_p} p(i)x(n-i)$$

where p(i) is a target response and x(n-i) represents the previously transmitted levels.

6. The transmitter of claim 5 wherein the target response, p(n), and the prefilter response, g(n), are derived from the predetermined response, c(n), of the analog channel.

7. The transmitter of claim 5 wherein the constellation point selector transmits the levels, x(n), according to the following function:

$$x(n) = y(n) - \sum_{i=1}^{N_p} p(i)x(n-i)$$

where y(n) are the constellation points.

8. The transmitter of claim 7 wherein the constellation point selector selects the constellation point in each equivalence class which minimizes the transmit power of the transmitter by selecting the constellation point, y(n), which produces the smallest value for x(n).

9. A method for providing a precoded sequence of analog levels over an analog channel to a quantization device, comprising:

mapping data bits to be transmitted to a sequence of equivalence classes, wherein each equivalence class contains one or more constellation points;

selecting a constellation point in each equivalence class to represent the data bits to be transmitted; and,

transmitting a level that produces the selected constellation point to an input of the quantization device.

10. The transmitter of claim 9 wherein the step of selecting a constellation point includes filtering the previously selected constellation points with a filter device and selecting the constellation points based on the output of the filter device.

11. The method of claim 10 further including filtering the level transmitted with a prefilter having a predefined filter response, g(n).

12. The method of claim 11 wherein the response of the filter device is:

$$\sum_{i=1}^{N_p} p(i)x(n-i)$$

where p(i) is a target response and x(n-i) represents the previously transmitted levels.

13. The method of claim 12 wherein the target response, p(n), and the prefilter response, g(n), are derived from the predetermined response, c(n), of the analog channel.

14. The method of claim 12 wherein step of transmitting includes transmitting the levels, x(n), according to the following function:

$$x(n) = y(n) - \sum_{i=1}^{N_p} p(i)x(n-i)$$

where y(n) are the constellation points.

15. The method of claim 14 wherein the step of selecting includes selecting the constellation point in each equivalence class which minimizes the transmit power of the transmitter by selecting the constellation point, y(n), which produces the smallest value for x(n).

16. A computer useable medium having computer readable program code means embodied therein to function as a precoder for transmitting a precoded sequence of analog levels over an analog channel to a quantization device, comprising:

computer readable program code means for mapping data bits to be transmitted to a sequence of equivalence classes, wherein each equivalence class contains one or more constellation points;

computer readable program code means for selecting a constellation point in each equivalence class to represent the data bits to be transmitted; and,

computer readable program code means for transmitting a level that produces the selected constellation point to an input of the quantization device.

17. The computer useable medium of claim 16 wherein the computer readable program code means for selecting a constellation point includes computer readable program code means for filtering the previously selected constellation points with a filter device and selecting the constellation points based on the output of the filter device.

18. The computer useable medium of claim 17 further including computer readable program code means for filtering the level transmitted with a prefilter having a predefined filter response, g(n).

19. The computer useable medium of claim 18 wherein the response of the filter device is:

$$\sum_{i=1}^{N_p} p(i)x(n-i)$$

where p(i) is a target response and x(n-i) represents the previously transmitted levels.

20. The computer useable medium of claim 19 further including computer readable program code means for deriving the target response, p(n), and the prefilter response, g(n), from the predetermined response, c(n), of the analog channel.

21. The computer useable medium of claim 19 wherein the computer readable program code means for transmitting includes computer readable program code means for transmitting the levels, x(n), according to the following function:

$$x(n) = y(n) - \sum_{i=1}^{N_p} p(i)x(n-i)$$

where y(n) are the constellation points.

22. The computer useable medium of claim 21 wherein the computer readable program code means for selecting includes computer readable program code means for selecting the constellation point in each equivalence class which minimizes the transmit power of the transmitter by selecting the constellation point, y(n), which produces the smallest value for x(n).

23. A computer data signal embodied in a carrier wave, wherein embodied in the computer data signal are computer readable program code means to function as a precoder for transmitting a precoded sequences of analog levels over an analog channel to a quantization device, wherein the analog channel modifies the transmitted analog levels, comprising:

computer readable program code means for mapping data bits to be transmitted to a sequence of equivalence classes, wherein each equivalence class contains one or more constellation points;

computer readable program code means for selecting a constellation point in each equivalence class to represent the data bits to be transmitted; and,

computer readable program code means for transmitting a level that produces the selected constellation point to an input of the quantization device.

24. The computer data signal of claim 23 wherein the computer readable program code means for selecting a constellation point includes computer readable program code means for filtering the previously selected constellation points with a filter device and selecting the constellation points based on the output of the filter device.

25. The computer data signal of claim 24 further including computer readable program code means for filtering the level transmitted with a prefilter having a predefined filter response, g(n).

26. The computer data signal of claim 25 wherein the response of the filter device is:

$$\sum_{i=1}^{N_p} p(i)x(n-i)$$

where p(i) is a target response and x(n-i) represents the previously transmitted levels.

27. The computer data signal of claim 26 further including computer readable program code means for deriving the target response, p(n), and the prefilter response, g(n), from the predetermined response, c(n), of the analog channel.

28. The computer data signal of claim 26 wherein the computer readable program code means for transmitting includes computer readable program code means for transmitting the levels, x(n), according to the following function:

$$x(n) = y(n) - \sum_{i=1}^{N_p} p(i)x(n-i)$$

where y(n) are the constellation points.

29. The computer data signal of claim 28 wherein the computer readable program code means for selecting includes computer readable program code means for select-

ing the constellation point in each equivalence class which minimizes the transmit power of the transmitter by selecting the constellation point, y(n), which produces the smallest value for x(n).

30. In an Analog pulse code modulation (PCM) modem adapted for upstream PCM data transmission to a digital PCM modem, a precoder for preceding a sequence of analog levels transmitted over an analog channel to a quantization device, comprising:

a mapping device for mapping data bits to be transmitted to a sequence of equivalence classes, wherein each equivalence class contains one or more constellation points; and,

a constellation point selector interconnected to the mapping device which selects a constellation point in each equivalence class to represent the data bits to be transmitted and which transmits an analog level that produces the selected constellation point at an input of the quantization device.

31. In an analog pulse code modulation (PCM) modem adapted for PCM data transmission to another analog PCM modem, a transmitter for precoding a sequence of analog levels transmitted over an analog channel to a quantization device, wherein the analog channel modifies the transmitted analog levels, the transmitter comprising:

a mapping device for mapping data bits to be transmitted to a sequence of equivalence classes, wherein each equivalence class contains one or more constellation points; and

a constellation point selector interconnected to the mapping device which selects a constellation point in each equivalence class to represent the data bits to be transmitted and which transmits an analog level that produces the selected constellation point at an input to the quantization device.

* * * * *

UNITED STATES PATENT AND TRADEMARK OFFICE

CERTIFICATE OF CORRECTION

PATENT NO. : 6,198,776 B1
DATED : March 6, 2001
INVENTOR(S) : Eyuboglu, Vedat M. et al.

Page 1 of 1

It is certified that error appears in the above-identified patent and that said Letters Patent is hereby corrected as shown below:

Abstract,
The 1st line which reads: "A device and method for preceding data signals for pulse.... delete "preceding" and insert -- precoding -- such that the line reads, "A device and method for precoding data signals for pulse"

Signed and Sealed this

Sixteenth Day of October, 2001

Attest:

Nicholas P. Godici

NICHOLAS P. GODICI

Attesting Officer

Acting Director of the United States Patent and Trademark Office

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JS 44 (Rev. 3/99)

CIVIL COVER SHEET

The JS-44 civil cover sheet and the information contained herein neither replace nor supplement the filing and service of pleadings or other papers as required by law, except as provided by local rules of court. This form, approved by the Judicial Conference of the United States in September 1974, is required for the use of the Clerk of Court for the purpose of initiating the civil docket sheet. (SEE INSTRUCTIONS ON THE REVERSE OF THE FORM.)

I. (a) PLAINTIFFS

CIF LICENSING, LLC, d/b/a GE LICENSING

(b) County of Residence of First Listed Plaintiff ____________
(EXCEPT IN U.S. PLAINTIFF CASES)

(c) Attorney's (Firm Name, Address, and Telephone Number)

Potter Anderson & Corroon LLP
P.O. Box 951
Wilmington, DE 19899
(302) 984-6000

DEFENDANTS

AGERE SYSTEMS INC.

County of Residence of First Listed Defendant ____________
(IN U.S. PLAINTIFF CASES ONLY)
NOTE: IN LAND CONDEMNATION CASES, USE THE LOCATION OF THE LAND INVOLVED.

Attorneys (If Known)

II. BASIS OF JURISDICTION (Place an "X" in One Box Only)

- [ ] 1 U.S. Government Plaintiff
- [x] 3 Federal Question (U.S. Government Not a Party)
- [ ] 2 U.S. Government Defendant
- [ ] 4 Diversity (Indicate Citizenship of Parties in Item III)

III. CITIZENSHIP OF PRINCIPAL PARTIES (Place an "X" in One Box for Plaintiff and One Box for Defendant)

(For Diversity Cases Only)

	PTF	DEF		PTF	DEF
Citizen of This State	1	1	Incorporated *or* Principal Place of Business In This State	4	4
Citizen of Another State	2	2	Incorporated *and* Principal Place of Business In Another State	5	5
Citizen or Subject of a Foreign Country	3	3	Foreign Nation	6	6

IV. NATURE OF SUIT (Place an "X" in One Box Only)

CONTRACT
- 110 Insurance
- 120 Marine
- 130 Miller Act
- 140 Negotiable Instrument
- 150 Recovery of Overpayment & Enforcement of Judgment
- 151 Medicare Act
- 152 Recovery of Defaulted Student Loans (Excl. Veterans)
- 153 Recovery of Overpayment of Veteran's Benefits
- 160 Stockholders' Suits
- 190 Other Contract
- 195 Contract Product Liability

REAL PROPERTY
- 210 Land Condemnation
- 220 Foreclosure
- 230 Rent Lease & Ejectment
- 240 Torts to Land
- 245 Tort Product Liability
- 290 All Other Real Property

TORTS

PERSONAL INJURY
- 310 Airplane
- 315 Airplane Product Liability
- 320 Assault, Libel & Slander
- 330 Federal Employers' Liability
- 340 Marine
- 345 Marine Product Liability
- 350 Motor Vehicle
- 355 Motor Vehicle Product Liability
- 360 Other Personal Injury

PERSONAL INJURY
- 362 Personal Injury— Med. Malpractice
- 365 Personal Injury — Product Liability
- 368 Asbestos Personal Injury Product Liability

PERSONAL PROPERTY
- 370 Other Fraud
- 371 Truth in Lending
- 380 Other Personal Property Damage
- 385 Property Damage Product Liability

CIVIL RIGHTS
- 441 Voting
- 442 Employment
- 443 Housing/ Accommodations
- 444 Welfare
- 440 Other Civil Rights

PRISONER PETITIONS
- 510 Motions to Vacate Sentence
- Habeas Corpus:
- 530 General
- 535 Death Penalty
- 540 Mandamus & Other
- 550 Civil Rights
- 555 Prison Condition

FORFEITURE/PENALTY
- 610 Agriculture
- 620 Other Food & Drug
- 625 Drug Related Seizure of Property 21 USC
- 630 Liquor Laws
- 640 R.R. & Truck
- 650 Airline Regs.
- 660 Occupational Safety/Health
- 690 Other

LABOR
- 710 Fair Labor Standards Act
- 720 Labor/Mgmt. Relations
- 730 Labor/Mgmt.Reporting & Disclosure Act
- 740 Railway Labor Act
- 790 Other Labor Litigation
- 791 Empl. Ret. Inc. Security Act

BANKRUPTCY
- 422 Appeal 28 USC 158
- 423 Withdrawal 28 USC 157

PROPERTY RIGHTS
- 820 Copyrights
- [x] 830 Patent
- 840 Trademark

SOCIAL SECURITY
- 861 HIA (1395ff)
- 862 Black Lung (923)
- 863 DIWC/DIWW (405(g))
- 864 SSID Title XVI
- 865 RSI (405(g))

FEDERAL TAX SUITS
- 870 Taxes (U.S. Plaintiff or Defendant)
- 871 IRS—Third Party 26 USC 7609

OTHER STATUTES
- 400 State Reapportionment
- 410 Antitrust
- 430 Banks and Banking
- 450 Commerce/ICC Rates/etc.
- 460 Deportation
- 470 Racketeer Influenced and Corrupt Organizations
- 810 Selective Service
- 850 Securities/Commodities/ Exchange
- 875 Customer Challenge 12 USC 3410
- 891 Agricultural Acts
- 892 Economic Stabilization Act
- 893 Environmental Matters
- 894 Energy Allocation Act
- 895 Freedom of Information Act
- 900 Appeal of Fee Determination Under Equal Access to Justice
- 950 Constitutionality of State Statutes
- 890 Other Statutory Actions

V. ORIGIN (PLACE AN "X" IN ONE BOX ONLY)

- [x] 1 Original Proceeding
- [ ] 2 Removed from State Court
- [ ] 3 Remanded from Appellate Court
- [ ] 4 Reinstated or Reopened
- [ ] 5 Transferred from another district (specify)
- [ ] 6 Multidistrict Litigation
- [ ] 7 Appeal to District Judge from Magistrate Judgment

VI. CAUSE OF ACTION

(Cite the U.S. Civil Statute under which you are filing and write brief statement of cause. Do not cite jurisdictional statutes unless diversity.)

This is an action for patent infringement arising out of the patent laws of the United States of America, 35 U.S.C., Section 101, et seq.

VII. REQUESTED IN COMPLAINT:

- [ ] CHECK IF THIS IS A **CLASS ACTION** UNDER F.R.C.P. 23

DEMAND $ ____________

CHECK YES only if demanded in complaint:
JURY DEMAND: [x] Yes [ ] No

VIII. RELATED CASE(S) IF ANY

(See instructions): JUDGE ____________ DOCKET NUMBER ____________

DATE 3-23-07

SIGNATURE OF ATTORNEY OF RECORD

FOR OFFICE USE ONLY

RECEIPT # ________ AMOUNT ________ APPLYING IFP ________ JUDGE ________ MAG. JUDGE ________

JS 44 Reverse (Rev. 12/96)

INSTRUCTIONS FOR ATTORNEYS COMPLETING CIVIL COVER SHEET FORM JS-44

Authority For Civil Cover Sheet

The JS-44 civil cover sheet and the information contained herein neither replaces nor supplements the filings and service of pleading or other papers as required by law, except as provided by local rules of court. This form, approved by the Judicial Conference of the United States in September 1974, is required for the use of the Clerk of Court for the purpose of initiating the civil docket sheet. Consequently, a civil cover sheet is submitted to the Clerk of Court for each civil complaint filed. The attorney filing a case should complete the form as follows:

I. (a) Plaintiffs-Defendants. Enter names (last, first, middle initial) of plaintiff and defendant. If the plaintiff or defendant is a government agency, use only the full name or standard abbreviations. If the plaintiff or defendant is an official within a government agency, identify first the agency and then the official, giving both name and title.

(b.) County of Residence. For each civil case filed, except U.S. plaintiff cases, enter the name of the county where the first listed plaintiff resides at the time of filing. In U.S. plaintiff cases, enter the name of the county in which the first listed defendant resides at the time of filing. (NOTE: In land condemnation cases, the county of residence of the "defendant" is the location of the tract of land involved.)

(c) Attorneys. Enter the firm name, address, telephone number, and attorney of record. If there are several attorneys, list them on an attachment, noting in this section "(see attachment)".

II. Jurisdiction. The basis of jurisdiction is set forth under Rule 8(a), F.R.C.P., which requires that jurisdictions be shown in pleadings. Place an "X" in one of the boxes. If there is more than one basis of jurisdiction, precedence is given in the order shown below.

United States plaintiff. (1) Jurisdiction based on 28 U.S.C. 1345 and 1348. Suits by agencies and officers of the United States, are included here.

United States defendant. (2) When the plaintiff is suing the United States, its officers or agencies, place an "X" in this box.

Federal question. (3) This refers to suits under 28 U.S.C. 1331, where jurisdiction arises under the Constitution of the United States, an amendment to the Constitution, an act of Congress or a treaty of the United States. In cases where the U.S. is a party, the U.S. plaintiff or defendant code takes precedence, and box 1 or 2 should be marked.

Diversity of citizenship. (4) This refers to suits under 28 U.S.C. 1332, where parties are citizens of different states. When Box 4 is checked, the citizenship of the different parties must be checked. (See Section III below; federal question actions take precedence over diversity cases.)

III. Residence (citizenship) of Principal Parties. This section of the JS-44 is to be completed if diversity of citizenship was indicated above. Mark this section for each principal party.

IV. Nature of Suit. Place an "X" in the appropriate box. If the nature of suit cannot be determined, be sure the cause of action, in Section IV below, is sufficient to enable the deputy clerk or the statistical clerks in the Administrative Office to determine the nature of suit. If the cause fits more than one nature of suit, select the most definitive.

V. Origin. Place an "X" in one of the seven boxes.

Original Proceedings. (1) Cases which originate in the United States district courts.

Removed from State Court. (2) Proceedings initiated in state courts may be removed to the district courts under Title 28 U.S.C., Section 1441. When the petition for removal is granted, check this box.

Remanded from Appellate Court. (3) Check this box for cases remanded to the district court for further action. Use the date of remand as the filing date.

Reinstated or Reopened. (4) Check this box for cases reinstated or reopened in the district court. Use the reopening date as the filing date.

Transferred from Another District. (5) For cases transferred under Title 28 U.S.C. Section 1404(a) Do not use this for within district transfers or multidistrict litigation transfers.

Multidistrict Litigation. (6) Check this box when a multidistrict case is transferred into the district under authority of Title 28 U.S.C. Section 1407. When this box is checked, do not check (5) above.

Appeal to District Judge from Magistrate Judgment. (7) Check this box for an appeal from a magistrate judge's decision.

VI. Cause of Action. Report the civil statute directly related to the cause of action and give a brief description of the cause.

VII. Requested in Complaint. Class Action. Place an "X" in this box if you are filing a class action under Rule 23, F.R.Cv.P.

Demand. In this space enter the dollar amount (in thousands of dollars) being demanded or indicate other demand such as a preliminary injunction.

Jury Demand. Check the appropriate box to indicate whether or not a jury is being demanded.

VIII. Related Cases. This section of the JS-44 is used to reference related pending cases if any. If there are related pending cases, insert the docket numbers and the corresponding judge names for such cases.

Date and Attorney Signature. Date and sign the civil cover sheet.

AO FORM 85 RECEIPT (REV. 9/04)

United States District Court for the District of Delaware

Civil Action No. 07-170

ACKNOWLEDGMENT OF RECEIPT FOR AO FORM 85

NOTICE OF AVAILABILITY OF A UNITED STATES MAGISTRATE JUDGE TO EXERCISE JURISDICTION

I HEREBY ACKNOWLEDGE RECEIPT OF 2 COPIES OF AO FORM 85.

3/23/07
(Date forms issued)

x Dustin Frohlich
(Signature of Party or their Representative)

Dustin Frohlich
(Printed name of Party or their Representative)

Note: Completed receipt will be filed in the Civil Action